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Dirk Jaeger, Christoph Schaaf
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High Performance Data Transmission on Cable
– Technology, Implementation, Networks –

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DVB-C2:
High Performance Data Transmission on Cable
– Technology, Implementation, Networks –

The publication contains contributions of the following experts:

Nr.	Title of Chapter	Author	Company
1	Introduction	PETER SIEBERT	DVB Project
2	Commercial Requirements	BART BRUSSE	Contest Consultancy
3	Architecture	CHRISTOPH SCHAAF	Kabel Deutschland
4	Input Processing	FRANK HERMANN	Panasonic
5	FEC Processing, Bit-interleaving & Mapping	Marten Kabutz	Thomson (Technicolor)
6	PLPs and Data Slices	JÖRG ROBERT	Institut für Nachrichten-technik, TUBS
7	Time Interleaving	WOO SUK KO	LG
8	Frequency Interleaving	SAM ATUNGSIRI	SONY
9	Usage of Pilots	TAKU YAMAGATA	Imagination Technologies
10	Framing	LOTHAR STADELMEIER	SONY
11	L1 & L2 Signalling and Preamble Structure	SANG-CHUL MOON	LG
12	Preamble Signal Protection Mechanisms	SEHO MYUNG	Samsung
13	OFDM Generation, RF Characteristics, Receiver Implementation	MICHAEL HEISENBERG	Kathrein
14	Simulation and Validation	PHILIPP HASSE	Institut für Nachrichten-technik, TUBS
15	System Aspects and Performance	DIRK JAEGER	Institut für Nachrichten-technik, TUBS
16	Network Planning	JAN DE NIJS	TNO
17	Head-end Architectures	ERNST FREESE	BLANKOM Digital
18	Implementation Scenarios	PETER FLÖTGEN	Unitymedia

Preface from the Editors

With the introduction of digital broadcasting some 15 years ago, the era of converging electronic media services entered the cable market. DVB-C was the first digital transmission standard applied in broadband cable networks. The 64-QAM constellation was commonly utilized as state of the art modulation technique supporting the transmission of 6 bit per symbol while allowing cable operators to transmit 6 digital TV channels instead of a single analogue TV channel through a physical 8 MHz cable channel. This achievement was a major step increasing the capacity of the networks.

The idea of operating a single digital cable system for full service provision triggered a development aiming at an integration of DVB-C in the Data Over Cable Service Interface Specification (DOCSIS®). The so-called EuroDOCSIS system has been using DVB-C for downstream transmission while adding upstream as well as MAC functionality. Through the application of this system in cable, DVB-C was used to not only provide television services but also telephony and fast Internet access. The commercial success of DVB-C is attested by the hundreds of millions products sold on world wide scale.

Over the last years the user demand for bandwidth has been growing permanently and estimations of various media intelligence companies are forecasting that this trend will continue to happen in the years to come. This hunger for bandwidth prompted the cable operators to further upgrade their networks which have been evolving from broadcast delivery media to broadband telecommunications infrastructures.

In 2007, a group of cable operators approached the DVB project with the request to create a new DVB standard for physical downstream transmission in cable networks. The standard should provide a significant increase in spectral efficiency as well as a flexible applicability of IP services. DVB experts groups were formed in 2008. The DVB Commercial Module prepared Commercial Requirements and the DVB Technical Module developed a technical specification for high performance data transmission on cable systems called DVB-C2.

The development started with a Study Mission giving a clear positive answer to the questions whether techniques have been evolved during the last 15 years, which legitimate the efforts to prepare the new standard and justify the assumption that a subsequent commercial introduction would become a commercial success. DVB's Call for Technology resulted in various responses which were combined to a final solution eventually agreed end of 2009.

At the time of publication of this book, the DVB-C2 standard is complete, approved, and published as DVB Blue Book A138 as well as European Standard EN 302 769. A first revised version of the standard incorporating minor modifications is underway. Implementers have started to develop DVB-C2 compliant equipment and first prototypes were exhibited at ANGA Cable 2010.

This book has the objective to provide a profound knowledge base on the DVB-C2 system gathered from DVB experts who have been actively involved in the development of the standard. This knowledge of the standardized techniques is accompanied by results of the work produced by the researchers engaged in the ReDeSign project which was co-funded by the European Union in the course of the EU's 7th Framework Programme.

The book takes the reader back to the origins of DVB-C2 by explaining the work process carried out by DVB to create the standard and introducing the Commercial Requirements which constitute the basis for the technical development. The major part of the book is dedicated to explain in great detail the individual techniques forming DVB-C2 and why they were selected. The third part presents additional information which gives DVB-C2 users practical data instrumental in producing standard compliant products and using these products in a real life environment, respectively.

Finally the book is seen as secondary literature accompanying the DVB-C2 standard (EN 302 769) itself as well as the DVB-C2 Implementation Guideline (TS 102 991).

Dirk Jaeger

Institut für Nachrichtentechnik,
TU Braunschweig
Chairman ReDeSign

Christoph Schaaf

Kabel Deutschland

Chairman DVB TM-C2

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1 Introduction

1.1 Projects and organizations

1.1.1 DVB Project and DVB-C2 development structure

The DVB Project was launched in 1993 with the objective to provide standards for digital television in Europe. Within just a few years, DVB developed the first generation of digital transmission specifications for satellite, cable, and terrestrial television. These standards proved to be extremely successful. It is estimated that over 500 Million DVB compliant devices have been deployed since 1994. To date several countries have introduced digital transmission and some have already switched off analogue terrestrial transmission altogether. Digital television based on DVB standards creates a win-win situation for everyone. DVB delivers more choice and convenience for the end user. For manufacturers, DVB specifications have created new markets and for network operators and broadcasters DVB has opened the door for new opportunities to reach their customers. They have also facilitated the launch of new and completely different business models.

Why has DVB been so successful? The key to the success of the DVB Project is its four basic principles:

- The involvement of all relevant market players
- Market driven
- Agreement by consensus
- Financed by member contribution

Firstly, DVB brings together the relevant players from the four major groups involved in broadcasting: broadcasters, network operators, regulators, and manufacturers. This mixture guarantees that DVB specifications, once they are published, are supported by all the relevant groups.

Secondly, DVB is a market driven body. At the planning stage of each new specification, the commercial requirements are discussed and agreed on by the Commercial Module. Once the commercial requirements have been approved, they become the basis for the technical work in the Technical Module. When a specification is finalised, the Commercial Module confirms that all their requirements have been considered.

Thirdly, decisions made within DVB are based on democratic principles. As a basic principle, commercial requirements and technical specifications must be agreed by consensus. Voting is only seen as a last resort to achieve a decision. This democratic principle helps accelerate the acceptance of DVB specifications by the market.

And finally, it is the DVB Members who contribute to the success by sending their experts to the meetings and support the specification work. The work of DVB is 100% funded by a modest membership fee and relies wholly on the voluntarily contribution of the members in the standardisation process. No government subsidies are involved in the work of DVB.

1.1.2 ReDeSign – a short introduction

ReDeSign is a research project (STREP) co-funded by the European Union under its 7^{th} Research Framework Program. Its 30 months lifespan began in January 2007 and ends in July 2010. The project brings together companies from the operator community, the vendor industry, and research organizations. During the first project year, three major cable operators namely the Belgium operator Telenet and ZON TV Cabo from Portugal were engaged in ReDeSign as well as ANGA Verband Deutscher Kabelnetzbetreiber, the German cable industry association. Furthermore, BLANKOM Digital (Germany) and VECTOR (Poland) represented the cable supplier industry while ALCATEL-Lucent contributed their knowhow in telecommunication systems. The research organizations are represented by the Dutch company TNO and Institut für Nachrichtentechnik of Technische Universität Braunschweig from Germany. The latter has coordinated the project.

The project's major objective has been to carry out research enabling cable operators to extend the lifetime of their current HFC infrastructures and to find the optimum strategy for the seamless migration of these infrastructures to the next generation HFC network architecture. In order to streamline the project's work to the requirements of the cable industry, a ReDeSign Operators' Forum was set up. At its two annual meetings, more than 20 cable operators discussed the project results and provided guidance for future work. Standards developing organizations were supported by the research results of the ReDeSign, which was an important objective for ReDeSign. Liaisons were established with CENELEC, DVB, and the European SCTE to foster close cooperation on matters of common interest. The support of DVB in developing DVB-C2 is a good example of such cooperation. ReDeSign provided valuable input to this activity, as explained in later chapters with examples. Further information on ReDeSign and the project's work results are available at www.ict-redesign.eu.

1.2 DVB 2.0 – second generation DVB standards

Figure 1 shows the development of DVB specifications since the project's inception in 1993. For cable operators, DVB-C, the first generation cable standard, created new business opportunities. DVB-C established the more efficient use of available bandwidth and significantly increased the number of programs that could be delivered to the home. In addition, together with the DOCSIS[1] return channel standard, cable became a bidirectional media. From that time onwards cable operators could offer high-speed Internet and voice services. With digital transmission, cable operators became players in telecommunication. Consequently, cable operators upgraded their networks and started delivering Triple Play services.

Ten years after the development of the first generation transmission standards, the market had significantly evolved:

- Industry and end-users were demanding High Definition Television (HDTV) services, and H.264, the latest video coding technology, provided a significant increase in bandwidth.

[1] DOCSIS® (Data Over Cable Service Interface Specification) is a cable modem certification project owned by the U.S. based Cable Television Laboratories Inc. (CableLabs®)

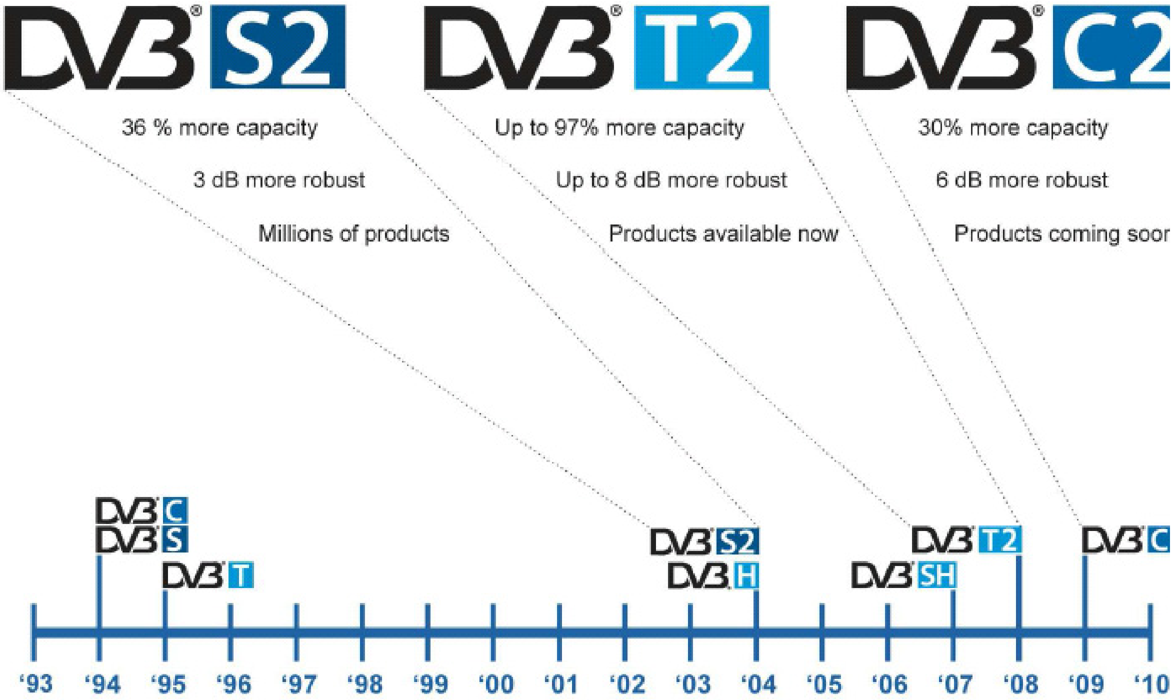


Figure 1: DVB development over more than 16 years

- Fuelled by Moore's Law, the latest System on Chips (SoC) can now support much more powerful algorithms and additional functionally at reduced cost.
- And finally, increased competition was forcing cable operators to look for new use cases and business models.

Taking into account all these changes and new requirements, DVB began to develop second generation broadcast specifications. DVB-C2, the latest in the DVB family, will successfully support cable operators for years to come by providing more capacity and additional flexibility. In this publication the main contributors to the DVB-C2 specification present the technologies on which DVB-C2 is based. One of which is the use OFDM, the underlying technology for DVB-T and DVB-T2. For instance OFDM allows a flexible bandwidths allocation. Instead of fixed frequency schemes of e.g. 6 or 8 MHz, operators now have the flexibility to offer any combination of bandwidths from 6 and 8 MHZ, respectively, up to some ten or even hundred MHz. Channels with greater bandwidths allow cable operators to deliver 100 Mbps to the home or office without using channel bundling technologies.

The development of the DVB-C2 specification took place in a close cooperation between DVB working groups and the ReDeSign Project, as shown in Figure 2. Members of the ReDeSign Project participated in defining the commercial requirements as well as the technical specifications.

For operators, DVB-C2 is coming at just the right point in time. Telephone and interactive data services over cable are growing at an exorbitant rate and take up evermore of the cable capacity on offer. In addition, there is the introduction of HDTV services requiring more bandwidth, the simulcast of standard definition, and the ongoing analogue services. Consequently, cable operators could soon run out of capacity. Furthermore, cable operators need improved delivery schemes to the home to compete with fibre and advanced DSL, supporting data rates in the range of 100 Mbps. DVB-C2 will offer this increased capacity as well as the new capabilities cable operators need to compete successfully in the years to come.

DVB-C2 is the latest member of the second generation broadcast family. The DVB Project has once more demonstrated that, driven by market requirements, the right technology is delivered at the right point in time. However, the work will not end there. DVB is committed to continue delivering world class standards, driving technology to new limits.

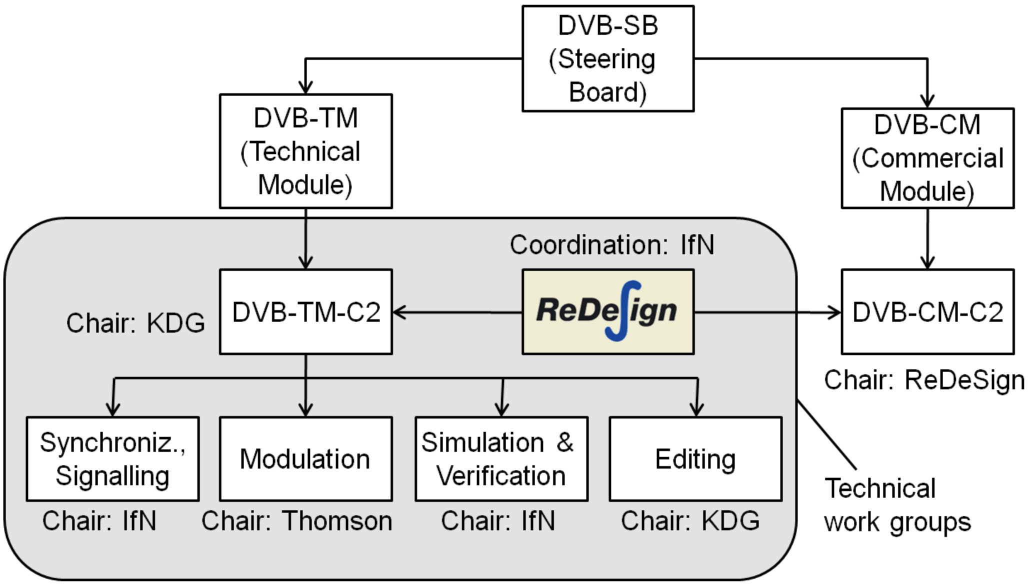


Figure 2: Block diagram of DVB-C2 development structure with technical work groups highlighted

2 Commercial Requirements

2.1 A new transmission technology for a new generation of digital services

Digital cable TV services started in various European markets in the late 20th century, enabled by DVB-C, the transmission technology developed by the DVB Project in 1994, and standardized by ETSI as EN 300 429. Since then, the technology was deployed in a large number of cable systems all over the world, enabling digital TV as well as broadband services for millions of customers.

However, since its first introduction, demand for more and more advanced digital TV and broadband services has been increasing every year, gradually pushing cable networks, even those that have been fully upgraded to 862 MHz HFC networks, to the limits of their capabilities in terms of capacity and flexibility. Cable operators have been seeking ways to offer HDTV packages, On-demand services, and a wider range of services based on more advanced interactive concepts. At the same time, experience gathered by the ReDeSign project through an extensive survey among cable operators has shown that analogue broadcasting via cable – despite aggressive switch-off scenarios in some countries – may continue for a considerable period, specifically in those markets where the majority of consumers rely on cable for the reception of television signals.

Whereas the 64-QAM and 256-QAM modulation schemes of the DVB-C technology at the time when the specification was developed, represented a reasonable compromise between spectrum efficiency, robustness against interference, and the operational complexity in a network environment loaded with analogue services, a new transmission standard was necessary in order to address the requirements generated by a new generation of more personalized digital TV and broadband services on cable, that is expected to dominate the second decade of the 21st century.

2.2 New services and a new specification

In today's changing market environment, it is evident that the number of digital channels in cable systems will increase considerably, specifically because many new services such as HDTV and VOD will demand much more bandwidth, while analogue TV cannot be terminated on short term. Moreover, experience teaches that service offerings need to be refreshed consistently, hence to remain competitive and flexible in these markets, mid to long term implementation of state of the art transmission technology will be required. This will also be necessary because many cable operators receive their signals via satellite or terrestrial networks, that are using – or will be using – higher modulation schemes, and in many situations it will be necessary to have the ability to retransmit an entire multiplex.

In addition, cable operators will increasingly seek opportunities to offer better services supporting more flexible service levels to residential and business customers. This will require tools offering more flexibility with respect to QoS and robustness, as well as with respect to the handling of IP based protocols, formats & codecs, and new technologies such as channel bonding. Moreover, an overall performance improvement will have to accompany the introduction of new services such as VOD and HDTV, in order to further

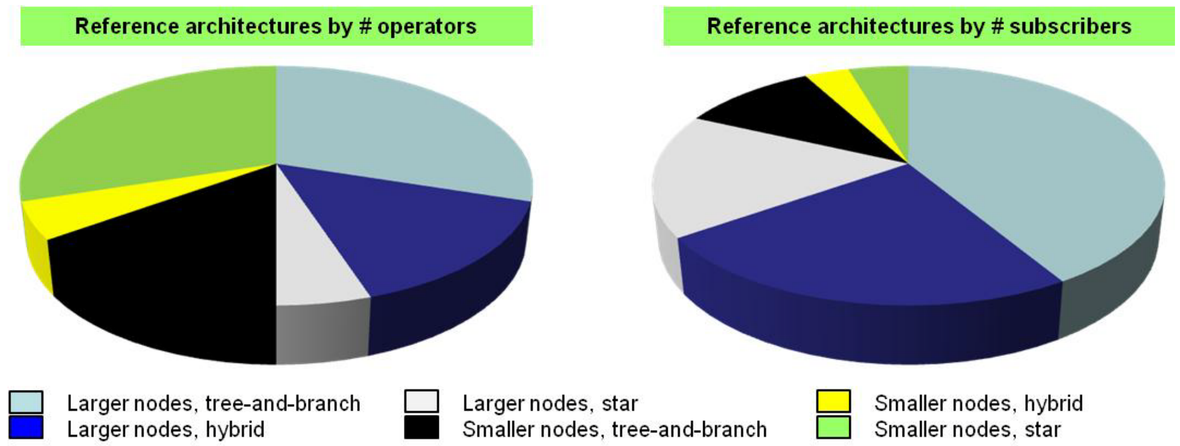


Figure 3: Distribution of reference architectures in the European cable landscape (Source: ReDeSign [3])

boost digital TV penetration, specifically when looking at time critical services such as on-line gaming an interactive TV.

The capability of cable operators to respond to the future market requirements as described above largely depends on: i) the network topologies and transmission technologies that are deployed today; ii) their expectations in terms of services take up and development, and; iii) the technologies that are available to upgrade their networks in a cost effective way.
When looking at the network topologies currently deployed by cable operators in Europe, research carried out by the ReDeSign project shows that generally speaking 6 types of cable networks can be found in Europe. These differentiate in terms of the type of topology used in the access network, and the size, in terms of homes passed, of the fibre node, bridging between the fibre and the coaxial part of an HFC network. Whereas at first sight different types of networks would appear to be distributed over cable operators relatively equally, a close look shows, that when taking the number of subscribers passed by a certain type of network as a criterion, the dominant cable network in Europe has a tree-and-branch or a hybrid (star/tree-and-branch) topology, as well as relatively large fibre nodes (see also Figure 3). Specifically the latter aspect not only means that downstream capacity shortages may occur relatively early when migration towards more personalized ('narrowcast') service concepts, but also that certain technologies may be less attractive than others, when aiming to create more capacity for these new service concepts.

When looking at cable operators' expectations with respect to the development of new services, research shows that new and more advanced digital TV and broadband services will gradually replace the analogue package as the backbone of the service portfolio during the second decade of the 21st century. Across Europe, the number of digital channels as well as the number of VOD users will strongly increase, with the number of digital receivers per household doubling, while the size of the analogue package will be downsize to an average of 20 channels.

When translating these expectations into concrete capacity requirements for existing as well as new services, analysis shows that an 'average' European cable network that does not have the full spectrum of 862 MHz available, will run into severe capacity shortages already in the 2011-2013 timeframe, whereas this will be the case in the 2014-2018 timeframe for networks that are fully upgraded. These results are depicted in the diagram in Figure 4

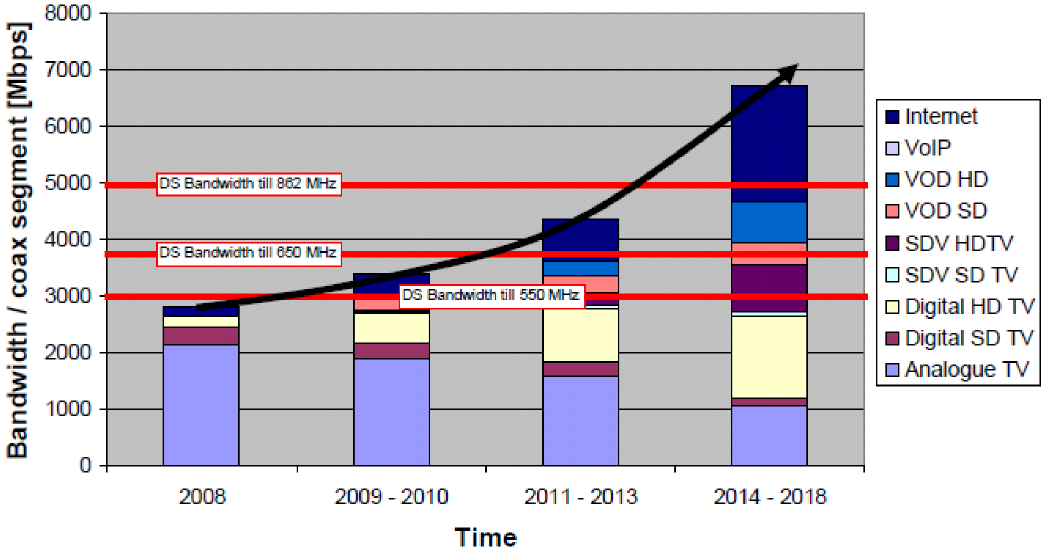


Figure 4: Average downstream bandwidth evolution for large fiber nodes using traditional transmission techniques (Source: ReDeSign [3])

A closer look at cable operators' expectations however reveals considerable differences between European regions, both with respect to the current size and shape of the analogue and digital services portfolio, as well as with respect to the way this will develop. Whereas operators in Central and Eastern European markets carry (and will be carrying on short to middle-term) relatively large analogue packages, these are considerably smaller in Western and Northern Europe. On the other hand, operators in Central Europe are already carrying large digital packages (and will increase these even more during the next years), whereas the digital package – at least on the short to middle term – will remain relatively moderate in size in the Western and Northern European markets.

These differences in expectations have considerable consequences for the development of capacity requirements in the different regional markets. When translating the figures into capacity requirements on the short, middle, and longer term, similar to what has been done in Figure 4 above, in Central and Eastern Europe, even fully upgraded cable networks will run into capacity shortages already in the 2011-2013 timeframe, whereas this will only be the case for cable networks in Western and Northern Europe on the longer term.

The findings of the ReDeSign project demonstrate that – as assumed by DVB when establishing the commercial basis for the new DVB-C2 specification – new and digital services such as HDTV, VOD and advanced broadband products will require an amount of downstream bandwidth that, under current circumstances, will not be available in the average European cable network (i.e. networks with relatively large fibre nodes and a tree-and-branch or hybrid star/tree-and-branch coaxial topology) on the mid to longer term. Moreover, in a very substantial amount of networks, this will even be the case on the short to middle-term. Consequently, solutions will be required that are capable of addressing these shortages (as well as other requirements originating from new services) on relatively short notice.

2.3 Evaluating concepts increasing capacity in cable networks

When evaluating several available upgrading technologies and their applicability to address expected capacity shortages as described above in the different European markets, their

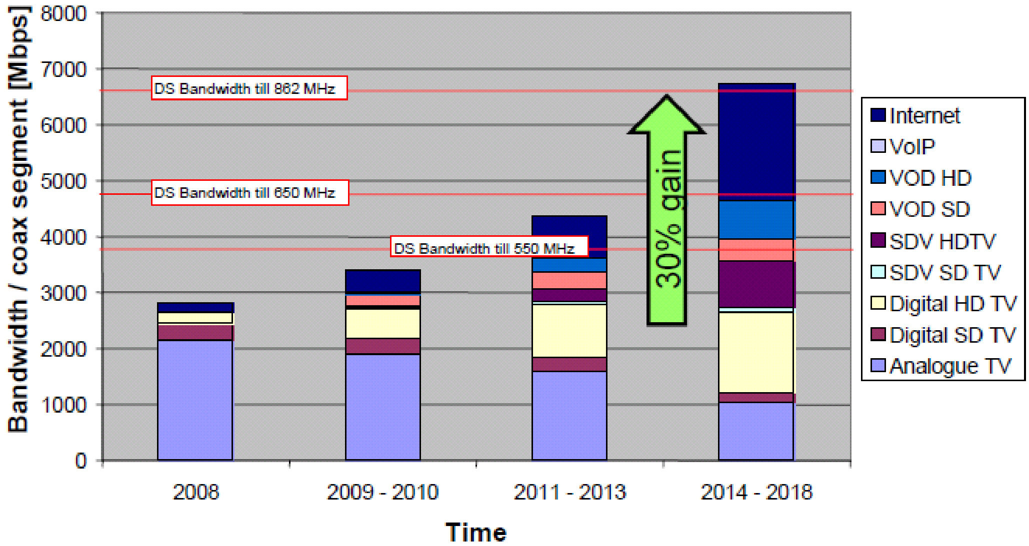


Figure 5: Average downstream bandwidth evolution for large fiber nodes with utilization of DVB-C2 (Source: ReDeSign [3])

character (can they be implemented cost effectively on short notice), as well as operators' preferences will have to be taken into account. Taking the latter criterion as a starting point, research carried out by ReDeSign shows that to most cable operators, splitting fibre nodes is the most relevant network upgrading technology, whereas only few are looking at extensions towards – or even beyond – 1 GHz. Both these concepts are however relatively expensive (when for example compared to concepts such as statistical multiplexing, QAM-sharing, or switching off analogue channels) as they involve 'physical' changes in the network. Moreover, both are strongly driven by the need to create more upstream, rather than downstream capacity.

A closer look at technologies that create more downstream capacity, either through more efficient usage of the transmission channel, or by more efficient usage of the available bandwidth, shows that better modulation codes are seen by many operators across virtually all markets as one of the more promising – and hence valuable – concepts. Moreover, contrary to for example Switched Digital Broadcast (where capacity gain is related to the size of the fibre nodes) or statistical multiplexing (which is often already implemented, either by operators or by broadcasters), the effect of better modulation codes is not dependent on other (technical) circumstances, and generates additional capacity gain on top of technologies already used. In this respect, an additional 30% capacity gain, generated by a new digital TV transmission system for cable, offering higher modulation profiles, could already restore most of the balance between network capacity and the development of bandwidth requirements as shown in Figure 5.

2.4 Commercial requirements

As pointed out earlier, research work of ReDeSign has demonstrated that new and advanced services will establish the framework for the provision of digital TV over cable in terms of capacity, robustness, and flexibility requirements at least for the upcoming 10 years. Furthermore the results have shown that it should be possible to implement a system addressing these requirements on relatively short notice (i.e. in the 2011-2013 timeframe). Taking all these investigation results into account, DVB defined the following commercial requirements for DVB-C2, the new transmission system for digital TV over cable.

Generally speaking, the technologies used shall aim to optimize the use of cable channels in state of the art cable networks, which includes enhanced flexibility and robustness, as well as maximum payload data capacity. Suitable techniques already in existence shall be adopted wherever possible, and DVB-C2 should reuse existing solutions for interfacing, coding and modulation wherever appropriate. Although the specification should not primarily aim to match DVB-S2 and/or DVB-T2, it should fully exploit its differentiating features to compete in the market of content delivery. Therefore downstream transmission technologies that maximally benefit from the availability of the return channel should be considered, although the specification itself shall not depend on the availability of a return channel.

A toolkit of system parameters shall be available to address consumer as well as business oriented applications, taking into account different performance levels of the CATV network. Therefore the specification shall allow service providers on cable networks to have individual quality of service targets, even for services within the same multiplex. Due account shall however be taken of the characteristics of cable networks currently deployed as well as of cost implications related to different devices for receivers and head-end equipment. The specification shall be transmission frequency neutral within typical cable frequency bands and the existing DVB-C standard shall not be modified nor shall changes to other specifications be required.

In terms of performance and efficiency requirements, DVB-C2 should efficiently support the migration from a mixed analogue/digital to a full digital network. It should be able to offer maximum performance/throughput in both kinds of networks. In this respect, DVB-C2 shall provide at least 30% more throughput in existing cable plants including in-house networks compared to 256-QAM (DVB-C). Moreover, DVB-C2 shall allow achieving the maximum benefit from statistical multiplexing.

Cable networks should be characterized and modelled on a global (e.g. US, Asia and Europe) level (including in-house network) and the best modulation/FEC schemes shall be selected taking into account a realistic cable channel model including the deployment of analogue PAL/SECAM/NTSC TV channels, the deployment of different digital signals (such as DVB, DOCSIS, Davic) and the associated signal back-off ratios to analogue signals as well as different noise types (white, burst, impulse), non-linearity and other interferences present in current and future networks.

The error performance of the system must be suitable for all types of services that may be carried. DVB-C2 shall provide a fully transparent link for Transport Stream, IP-packets, and other relevant protocols between the input of the modulator and the output of the demodulator. Seamless retransmission (e.g. from DVB-S2 to DVB-C2, or DVB-T2 to DVB-C2) should be fully supported. The time to tune a receiver from one service to another (zapping time) shall not be significantly increased due to the introduction of DVB-C2, i.e. for any change in RF channel, the DVB-C2 front-end shall deliver a quasi error free signal within 300 ms. Finally, the transmission system should be able to support low-power modes to maximally reduce power consumption in receivers according to the EU Code of Conduct on Energy Consumption.

DVB-C2 doesn’t need to be compatible with DVB-C, meaning a DVB-C receiver doesn’t have to be capable to process a DVB-C2 signal. The capability for a DVB-C2 receiver to include

DVB-C functionalities should be addressed as an optional requirement in the technical specification, so that – if required – DVB-C can be functionality included into DVB-C2 equipment. DVB-C2 transmissions shall not cause any need for changes to existing DVB-C receivers, and the new standard should be as insensitive as possible to typical characteristics of in-house networks using coaxial cable systems.

Finally, when looking at interactive systems requirements, the specification shall be available for consideration as an alternative downstream coding and modulation scheme for the DOCSIS systems currently using DVB-C in its EuroDOCSIS variant. It shall include techniques for improving the efficiency of carriage of IP data and shall allow cost effective integration of DVB-C2 into EdgeQAM solutions for modulation equipment. Also, DVB-C2 shall provide a low latency mode for those interactive services that require such a mode.

2.5 Final remarks

Taking the above mentioned commercial requirements as a starting point, during 2008 DVB developed its new transmission system for cable. During this process, technical choices and decisions were continuously taken against the reality of market developments, thus assuring that the new system would not only address the needs of all cable operators, but also that it would arrive in time to support the massive roll-out of new and advanced digital TV services.

Before its final approval, the technical DVB-C2 specification was measured against the commercial requirements that had been specified early 2008 at the start of the process. Conclusion of this evaluation was that the new system addresses all requirements adequately, and hence should offer a large number of operators in Europe as well as on a global level the capacity and flexibility they need to address the increasing demand for new services and more flexibility. Moreover, the fact that DVB-C2 is very close to DVB-T2 and DVB-S2, and includes features for addressing a large variety of different services, will make the standard attractive for implementation in virtually all market environments.

3 Architecture

The DVB-C standard was developed in 1994. At that time the chosen signal carrier modulation schemes with up to 256-QAM modulation was a revolutionary approach, as nobody had yet experiences with such complex modulation schemes in consumer type equipment. Adaptive equalisation was necessary in DVB-C demodulators and nobody knew the behaviour of mixed analogue and digital channel configurations in CATV networks.

From 1996 onwards cable operators started DVB-C deployment with 64-QAM (@ 38 Mbps) modulation. This choice was a good compromise between complexity and payload through put on the one side and sufficient headroom for a robust implementation in state of the art CATV networks on the other side. A further driving force was the fact, that the standard DVB-S parameters of DTH satellite systems (33 MHz, 27.5 MBaud, 3/4 FEC) allowed a seamless transition between satellite and cable, as DTH satellite links were often used to feed cable networks as well. After successful passing the DVB-C learning curve, more and more operators started to introduce 256-QAM (@ 50.7 Mbps) channels at 6 dB higher signal level in relation to 64-QAM DVB-C.

With the reduction of the analogue TV offering, cable operators gained intermodulation headroom in their networks. Network upgrades to so called Hybrid Coax Fibre (HFC) structures improved the performance of cable networks. DVB-C does not provide higher modulation constellation to utilize those available headrooms. Enhanced FEC schemes, as already integrated in other DVB-X2 transmission systems, were also promising in the cable application. Besides the MPEG Transport Stream more and more new protocols emerged for new services. So the three key requirements of DVB-C2 can be summarized as: Spectrum efficiency, flexibility and headroom for enhanced networks.

3.1 Single pipe versus multiple pipes and formats

DVB-C and all other first generation transmission systems were designed to carry one DVB Transport Stream. Although one Transport Stream usually contains several MPEG-encoded services, one key requirement for DVB-C2 to implement significantly more flexibility in terms of supporting multiple input signals and in terms of supporting more packetized and even continuous input formats, including the Internet Protocol. The flexibility of DVB-C2 allows the integration of different input signals in so called Physical Layer Pipes (PLPs) und to bundle such PLPs in so called Data Slices. Furthermore it is possible to extract identical data transmitted in several PLPs from these PLPs and insert them into so called common PLPs. By this means common data are transmitted only once and re-multiplexed to the relevant data PLP in the demodulator. DVB-C2 provides a very flexible multiplexing scheme, capable to support future complex services.

3.2 Single carrier versus orthogonal frequency division multiplexing (OFDM) modulation

More than 100 Million DVB-C based cable tuners, based on single carrier QAM-modulation, have been deployed worldwide over the last 15 years. So this modulation scheme was a strong candidate for DVB-C2 as well. In the course of the DVB Call for Technologies for DVB-C2 three proposals of OFDM based solution had been made as well. After a long technical debate and a detailed technical comparison of the pros and cons of the tow alternative modulation schemes, a unanimous decision was taken to choose for DVB-C2 the Orthogonal Frequency Division Multiplex (OFDM) modulation scheme for reasons of excellent spectrum efficiency and superb flexibility.

3.3 Low Density Parity Check (LDPC) code for FEC

The chosen Forward Error Correction (FEC) scheme is a combination of Low Density Parity Check (LDPC) code as the inner code and Bose Chaudhuri Hoquenghem (BCH) code as the outer code. The combination is both very powerful and efficient in relation to typical and relevant interference scenarios in cable networks. The excellent performance of the chosen FEC-scheme is the mayor reason for the significantly higher spectrum efficiency of DVB-C2. Those state of the art FEC codes are very complex. The LDPC-FEC processing part of the DVB-C2 decoder will require about half of the chip size. Bit interleaving is used to further optimize the FEC performance.

3.4 From 16-QAM to 4096-QAM constellations

The characteristics and performance figures of cable networks are covering a wide range from low cost Master Antenna TV (MATV) solutions to high quality professional HFC networks. Therefore DVB-C2 offers a fine granularity of solutions form very robust modes up to highest spectrum efficiency, mainly limited by cost constraints of receiver ADCs. Different FEC code rates and QAM-schemes allow the granularity of about 2 dB over the whole carrier-to-noise (CNR) range from 15 to 35 dB. Further higher modulation constellations may be introduced in the future in a backwards compatible way. At least there are already hooks available for future extensions of DVB-C2.

3.5 Fixed 8 MHz versus flexible bandwidth

Although DVB-C2 is perfectly in line with the European 8 MHz cable channel raster (as well as the 6 MHz US channel raster), one of the outstanding features of DVB-C2 is its flexibility in terms of bandwidth allocation. DVB-C2 allows increased spectrum efficiency and broader transmission signals entailing a higher gain for statistical multiplexing while maintaining the support for simple receivers with a fixed 8 MHz receiving window for Europe (6 MHz for the U.S.). For the implementation of future broadband tuner concepts, DVB-C2 opens more options for all kinds of broadband applications (up to 3.4 Gbps pipes using PLP bundling).

3.6 Constant coding and modulation (CCM) versus variable and adaptive coding and modulation (VCM and ACM)

DVB-C2 offers another dimension of flexibility. Up to now the coding schemes for cable transmission systems are fixed. With DVB-C2 the modulation parameters may vary over time and this variation may be even related to individual services within a transmitted DVB-C2 signal. The first option is to vary the robustness over time. This may be required for different Quality of Service (QoS) levels. However, it is also possible to adapt the performance of a DVB-C2 transmission to individual requirements of a customer by means of adaptive coding and modulation. The receiving conditions of an individual customer, reported via the integrated cable return channel to the transmitter, may be used to optimally adjust the robustness parameters of the DVB-C2 transmission according to the channel characteristics.

3.7 Physical Layer Pipes (PLP), Data Slices, and frames

The end user demand is permanently growing, In terms of both broadband access and video quality (e.g. HDTV). From a cable network operator's point of view, bigger pipes are required to transmit the requested services over cable networks in an efficient way. The big payload difference between narrowband and broadband services require flexible multiplexing schemes. DVB-C2 offers therefore a two stage multiplexing scheme. Different input signals, converted to so called Physical Layer Pipes (PLPs) are multiplexed to a Data Slice and different Data Slices are combined to a 'DVB-C2 Frame' in the second stage. In practice, DVB-C2 will of course be used to carry digital TV in simple broadcasting applications with a single MPEG Transport Stream converted to a single PLP included in a single Data Slice. In more complex services configurations, DVB-C2 will allow to structure the offering in different PLPs and Data Slices. Such assignment of dedicated PLPs to individual applications actually allows providing a service-related robustness as well as an adjustment of the payload capacity of PLPs or Data Slices, which may slightly vary over time.

3.8 Two dimensional interleaving in time and frequency domain

In relation to DVB-C the new standard offers both additional time and frequency interleaving, powerful tools to cope with critical interference scenarios in cable networks.

3.9 Signalling issues

The flexibility of DVB-C2 requires an appropriate signalling scheme, allowing a receiver a fast synchronisation and an easy access to all relevant parameters required to configure the demodulation and FEC processing of data related to the service chosen by the customer. The DVB-C2 systems distinguishes between Layer 1 signalling data carried in the preamble of a C2 system and Layer 2 signalling data which are part of Network Information Table (NIT´) of the DVB-SI information system.

3.10 Block modulator diagram

Figure 6 summarizes the four main building blocks of DVB-C2. The input processing multiplexes the input streams to the so called Physical Layer Pipes. The second processing stage does the FEC processing and the mapping to OFDM symbols. The second multiplexing stage in DVB-C2 is the Data Slice builder. Furthermore Data Slices, the preamble symbols, and the continual and scattered pilots are multiplexed to a DVB-C2 Frame in the third processing step. The fourth DVB-C2 building block is the OFDM conversion, which also includes the output processing.

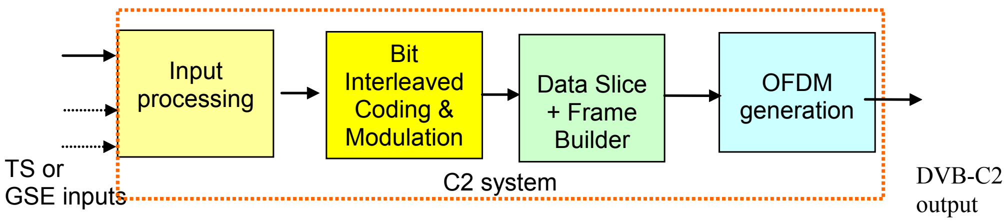


Figure 6: Overall DVB-C2 modulator block diagram

3.10.1 Modulator input processing

Figure 7 shows the main building blocks of the input processing part of a DVB-C2 modulator. Different input signals are synchronised and mapped into a baseband (BB) framing structure. Null packets may be deleted, if this increases the overall payload capacity. The BB headers carry the relevant parameters of the input processing related to the different BB frames.

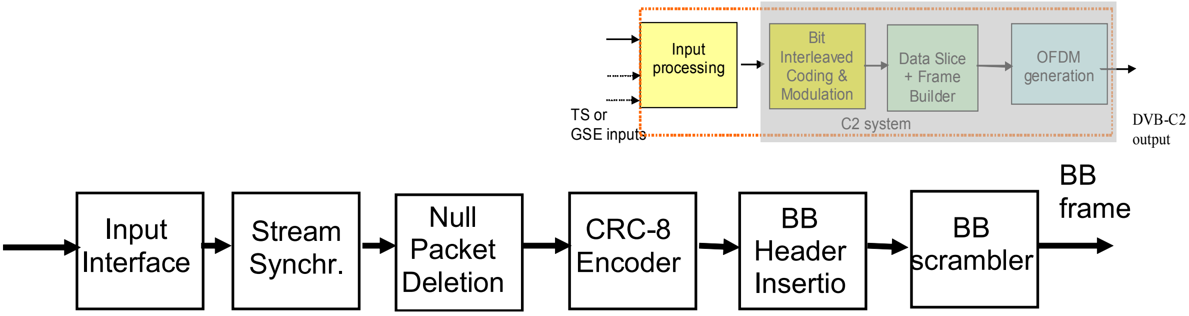


Figure 7: Building blocks of the input processing part

3.10.2 Bit interleaved FEC processing and mapping

Figure 8 shows the main building blocks of the FEC processing part of a DVB-C2 modulator. Baseband frames are extended by both BCH and LDPC FEC data. The bit stream is de-multiplexed and mapped to QAM-CELs. A FEC-Frame Header is added. DVB-C2 allows the mapping of both normal FECFrames (64,800 bit) e.g. for broadcasting applications and short FECFrames (16,200 bit) for time critical services.

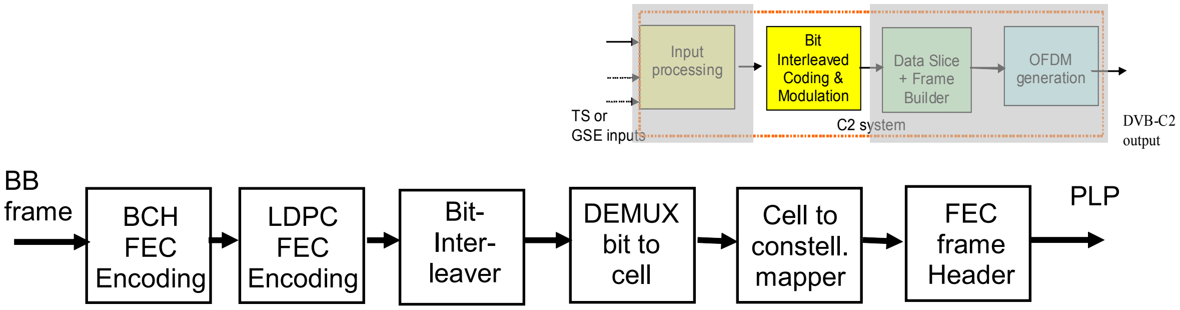


Figure 8: Building blocks of the FEC processing part

3.10.3 Data Slice and Frame Builder

Figure 9 and Figure 10 show the main building blocks of the Data Slice building part of a DVB-C2 modulator. In the first multiplexing stage different PLPs are multiplexed to a dada slice and both Time and Frequency Interleaving is applied to this Data Slice. In the second multiplexing stage different Data Slices, Pilots and the Preamble are multiplexed to a C2-Frame.

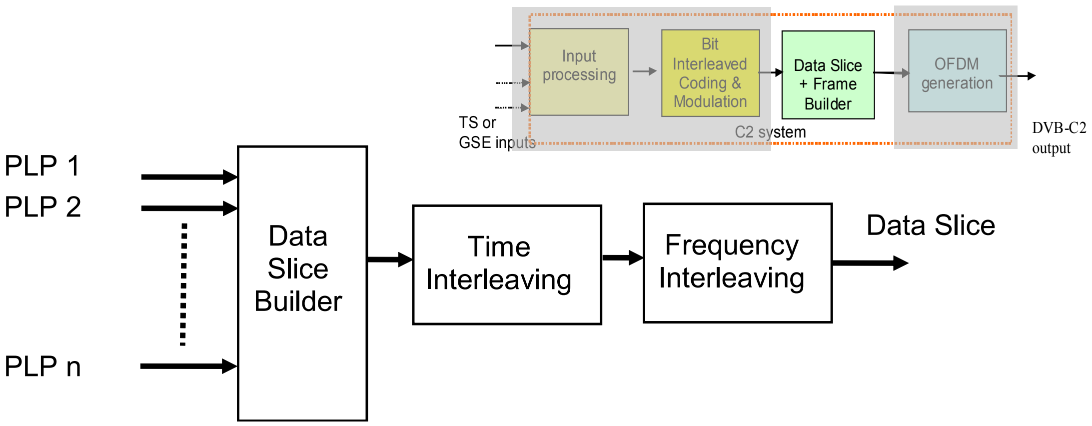


Figure 9: Building blocks of the Data Slice Building part

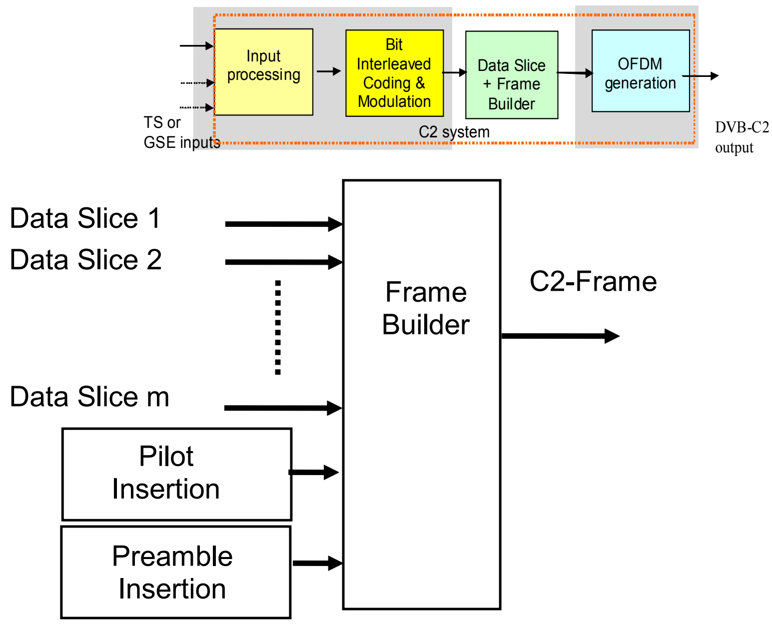


Figure 10: Building blocks of the Frame Building part

3.10.4 OFDM generation

Figure 11 shows the main building blocks of the OFDM generation unit of a DVB-C2 modulator. After the inverse FFT processing, the Guard Interval is added and an analogue to digital conversion is carried out. In the unlikely event of high PAPR, reserved tone symbols could be inserted.

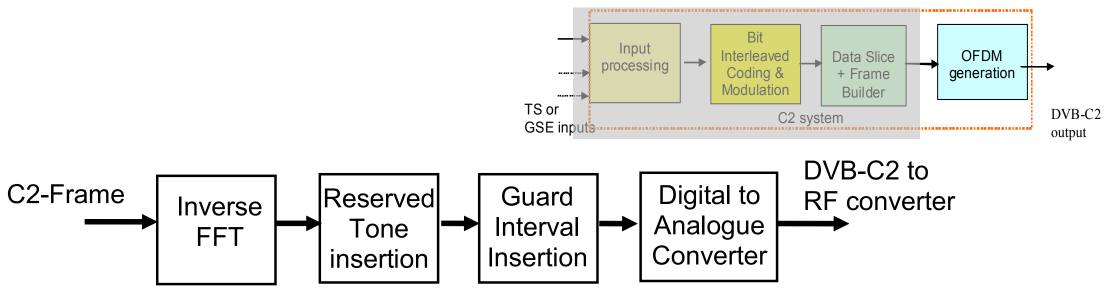


Figure 11: Building blocks of the OFDM generation part

4 Input Pre-processing and Processing

The input processing stage is the entry stage to the C2 physical layer. It performs the mode and the stream adaptation. These two adaptation steps consist of the input interfacing, input stream synchronization, null packet deletion, CRC-8 encoding and BBF header insertion on the one hand side and padding insertion as well as BBF scrambling on the other hand.

On top of the input processing sits the input pre-processing whose main functional element is the transport stream re-multiplexer and/or the traffic shaper for generic streams.

The position of these two entry stages is outlined with Figure 12 below, the detailed functional blocks with Figure 15, Figure 16, and Figure 20. The general processing tasks are described with the following sub-clauses.

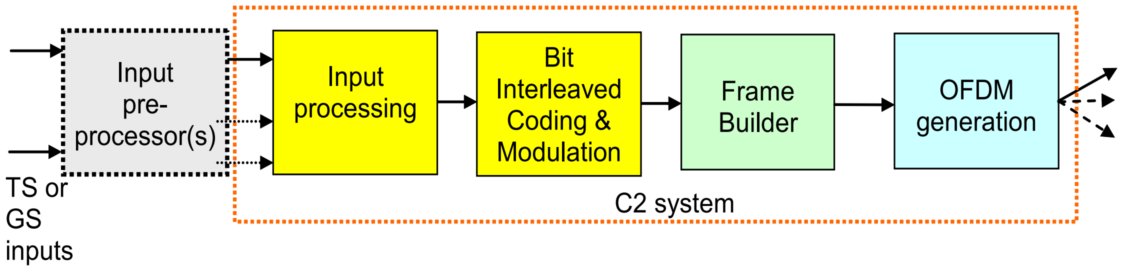


Figure 12: High level C2 system block diagram

Input processing and pre-processing for C2 is almost the same as for DVB-T2. There are four main differences:

(1) For T2 there is a compensating delay block per PLP that serves for compensating different time interleaving depths for the different PLPs. The presence of this buffer on the transmitter side saves memory on the receiver side that would otherwise be necessary for synchronization purposes. For C2 this stage is not required since the time interleaving depth used here is an individual setting for each Data Slice, but not for each PLP

(2) There is no L1-inband signalling in C2. This is due to the fact that pre-amble signalling of Layer 1 metadata is sufficient in the light of the limited disturbances on cable channels.

(3) In T2 there are quite strict rules applied regarding the content of the Common PLPs and the synchronization regarding the splitting and re-insertion of common data. This is because the target is to enable a 1:1 re-assembly of the original Transport Streams on the receiver side.
In the C2 case the re-assembly is done in a way that doesn't necessarily lead to the exact packet sequence of the original stream.

(4) Opposite to T2, C2 doesn't provide the option to interleave data of a single FECFrame over more than one C2 transmission frame.

4.1 Input pre-processing

Input streams to this stage are either Transport Streams (TSs) [1] or Generic Streams (GSs) [2]. The latter stream type can occur in three different forms, Generic Stream Encapsulation

(GSE), Generic Fixed-length Packetized Stream (GFPS)[2] and Generic Continuous Stream (GCS)[3]. The first GS variant might be the one that is employed most often nowadays.

These streams can be split by an optional re-multiplexer (for TSs) or traffic shaper (for GSs). Task of the re-multiplexing/traffic shaping is to define groups of streams, of which one carries information (i.e. EMMs, EPG data) that is commonly applicable to the remaining streams of the related group. The other streams belonging to the same group consist of the individual data like e.g. audio/video streams. As far as Transport Streams are concerned, they become partial streams after the split (TSPSs), i.e. they are not syntactically correct streams anymore when transmitted due to the separate transport of basic signalling information. That common information travels as a “Transport Stream Partial Stream Common” (TSPSC). The same concept might be applicable to Generic Streams as well. As far as the corresponding PLPs are concerned, they are identified as Data PLPs and Common PLPs. The receivers combine one partial stream (stemming from a Data PLP) with the related partial stream common (from Common PLP) and regenerate a syntactically correct stream again.

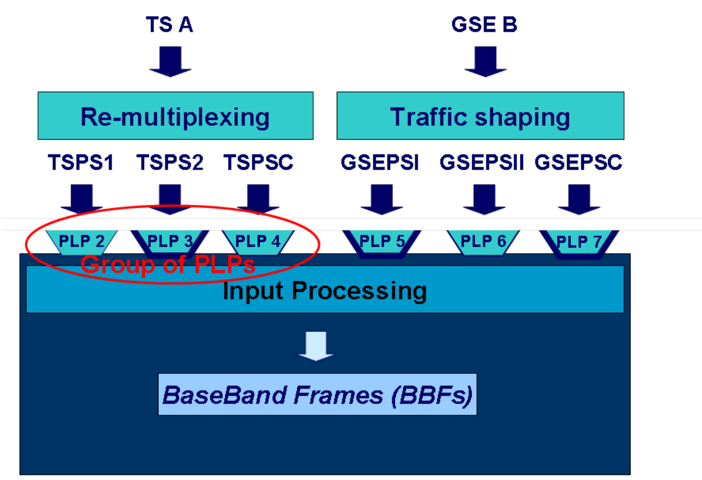


Figure 13: Re-multiplexing of TSs, traffic shaping of GSEs

4.2 Building the Common PLP

The C2 broadcasting system offers the option to transport services within individual PLPs. When each of those PLPs would carry all metadata applicable to its content and to the whole set of transported streams, a high degree of unnecessary redundancy would be generated. The solution to this is to carry common metadata (and other common data such as EMMs of the CA system) in a Common PLP. Basis for this is the grouping of PLPs that share that common data. A single multiplex can carry multiple PLP groups.

Opposite to T2, C2 is not heading for the re-assembly of exactly the same streams as those that reached the input of the re-multiplexer and/or traffic shaper (see Figure 13 above). Instead C2 envisages the re-insertion of the common data on the receiver side whenever there is space for it in the data (e.g. video/audio) stream. Therefore no such strict synchronization requirements as for T2 are applicable between Data and Common PLPs. Based on this more flexible approach, common data that is time-critical shall not be moved to the Common PLP. The positions in the Data PLPs where packets were moved from to the Common PLP shall be filled with null packets.

[2] GFPS:Generic Fixed-length Packetized Stream (GFPS); this form is retained for compatibility with DVB-S2, but it is expected that GSE would now be used instead
[3] GCS: Generic Continuous Stream, variable length packet stream where the modulator is not aware of the packet boundaries

The sheer dis- and re-assembly mechanism for the streams is conceptually the same as for T2:

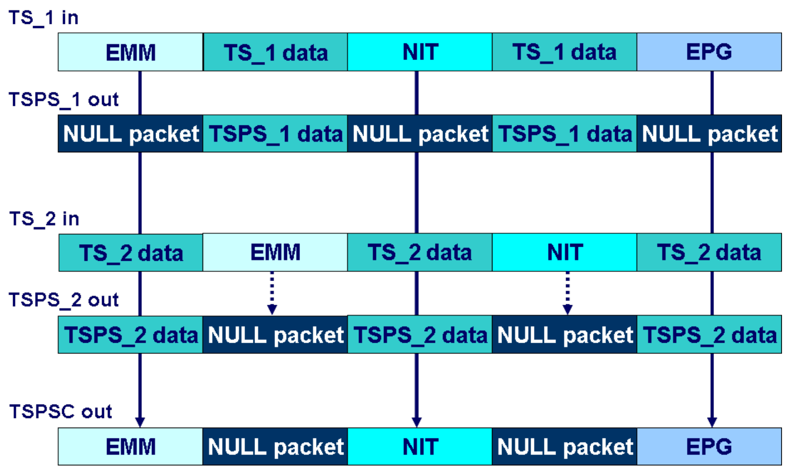


Figure 14: Building the Common PLP

4.3 Input processing

The input processing block consists of two stages, the mode adaptation and stream adaptation stage.

4.3.1 Mode adaptation

The mode adaptation looks slightly different for the single PLP case (Mode A) and the multiple PLP case (Mode B).

4.3.1.1 Single PLP case

In this simpler case where e.g. only a single transport stream is provided, the following processing blocks are applied:

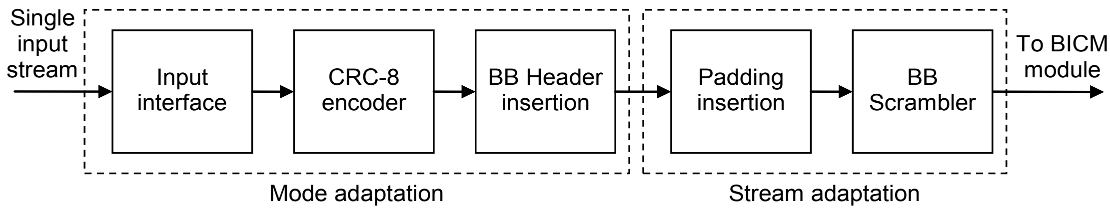


Figure 15: Input processing block

Input interface:
This block converts the format of the signal at the input to the format that is used internally for further processing.

CRC-8 encoder:
Error detection for the UPL fields of the user packets (after sync byte removal, if applicable) is enabled by means of the single CRC byte. The corresponding sync byte is appended after the related UP. Note that this detection is not 100% reliable.

Baseband frame header insertion:
This functional block configures the 10 byte header of each baseband frame (BBF) in line with the settings chosen by the provider and inserts it at the position of the most significant

bytes in the baseband frame, i.e. in front of the data field. The header describes the format of the data field. The important distinction between the two possible BBF modes, i.e. normal and high efficiency mode (NM, HEM), is provided with the least significant byte of the header (mode indication 'EXORed' with CRC-8 there). For further details on BBF headers see clause v below. Output to the stream adaptation block are unscrambled baseband frames filled with data and headers, but not consisting of padding so far in those cases where the data and the headers don't fill up the BBFs completely.

4.3.1.2 Multiple PLP case

In Mode B, two further functional blocks are part of the mode adaptation stage as outlined below.

<u>Input stream synchronizer</u>

Delays and packet jitter introduced by DVB-T2 modems may depend on the transmitted bit-rate and may change in time during bit and/or code rate switching. The "Input Stream Synchronizer" (see Figure 15 above) shall provide a mechanism to regenerate, in the receiver, the clock of the Transport Stream (or packetized Generic Stream) at the modulator Mode Adapter input, in order to guarantee end-to-end constant bit rates and delays. Table 1 gives the details of the coding of the ISSY field generated by the input stream synchronizer. When indicated by the ISSYI parameter in the MATYPE field, a counter shall be activated (22 bits), clocked by the modulator sampling rate (frequency Rs=1/T, where T is the elementary period). The Input Stream Synchronization field (ISSY, 2 or 3 bytes) shall then be transmitted according to clause v.

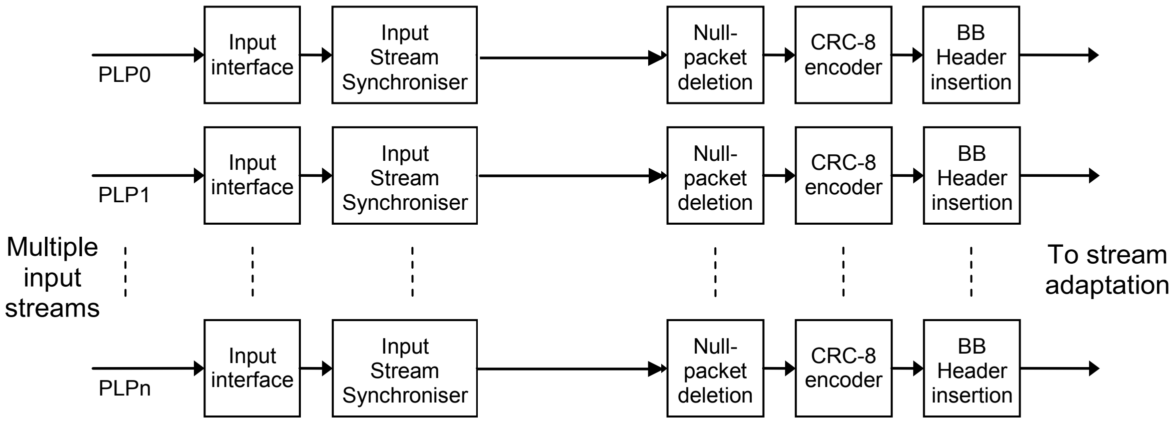


Figure 16: Mode adaptation for input mode B – multiple PLPs

An example generator/inserter for ISSY is illustrated with Figure 17 below.

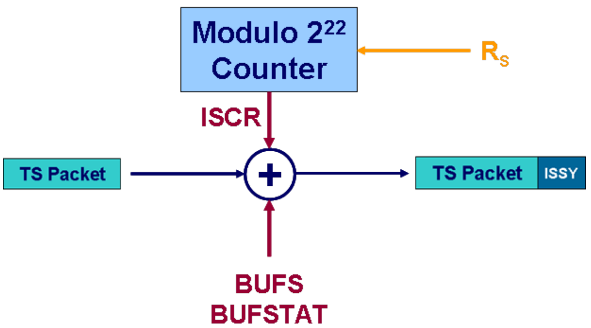


Figure 17: Input Stream Synchronizer block diagram for the TS case

First Byte					**Second Byte**	**Third Byte**
bit 7 (MSB)	bit 6	bit 5 and bit 4	bit 3 and bit 2	bit 1 and bit 0	bit 7 to bit-0	bit 7 bit 0
0 = ISCR short	MSB of ISCR short	next 6 bits of $ISCR_{short}$			next 8 bits of $ISCR_{short}$	not present
1	0 = ISCR long	6 MSBs of $ISCR_{long}$			next 8 bits of $ISCR_{long}$	next 8 bits of $ISCR_{long}$
1	1	00 = BUFS	BUFS unit 00 = bit 01 = Kbit 10 = Mbit 11 = 8Kbit	2 MSBs of BUFS	next 8 bits of BUFS	not present when $ISCR_{short}$ is used; else reserved for future use
1	1	10 = BUFSTAT	BUFSTAT unit 00 = bit 01 = Kbit 10 = Mbit 11 = BUFS/1024	2 MSBs of BUFSTAT	next 8 bits of BUFSTAT	not present when $ISCR_{short}$ is used; else reserved for future use
1	1	01 = TTO	4 MSBs of TTO_E		Bit 7:LSB of TTO_E Bit 6-Bit0: TTO_M	not present when $ISCR_{short}$ is used; else TTO_L
1	1	others = res. for future use	reserved for future use	Reserved for future use	Reserved for future use	not present when $ISCR_{short}$ is used; else reserved for future use

Table 1: ISSY parameter coding, 2 or 3 bytes

The ISSY parameters are:

- ISCR (short: 15 bits; long: 22 bits) (ISCR = Input Stream Time Reference), loaded with the LSBs of the counter content at the instant the relevant input packet is processed (at constant rate RIN). In the case of continuous streams the content of the counter is loaded when the MSB of the Data Field is processed.
  ISCR shall be transmitted for each PLP. In HEM, for BBFrames for which no UP begins in the Data Field, ISCR is not applicable and BUFS shall be sent instead.
  A PLP shall not change from short to long ISSY except at a reconfiguration. In HEM, $ISCR_{long}$ shall always be used.

- BUFS: This parameter indicates the maximum size of the requested receiver buffer to compensate delay variations of the related PLP. The maximum value in the C2 case is 2 Mbit.
- BUFSTAT: This variable is retained for compatibility with DVB-S2. It is not be used in DVB-C2.
- TTO: This parameter provides a mechanism to manage the de-jitter buffer in DVB-C2. TTO defines the time, in units of T, between the beginning of the P1 symbol of the first C2 frame carrying the relevant User Packet and the time at which the MSB of the User Packet should be output, for a receiver implementing the model defined in further down this sub-clause. This value may be used to set the receiver buffer status during reception start-up procedure, and to verify normal functioning in steady state. TTO shall be transmitted for each PLP.

The choice of the parameters of a DVB-C2 system and the use of TTO shall be such that, if a receiver obeys the TTO signalling and implements the model of buffer management defined in the clause below, the receiver's de-jitter buffer and time de-interleaver memory and frequency de-interleaver shall neither overflow nor underflow.

Receiver buffer model

In the cases where ISSY is used, the receiver buffer model outlined with Figure 18 below shall build the basis for the configuration of the transmitted signal.

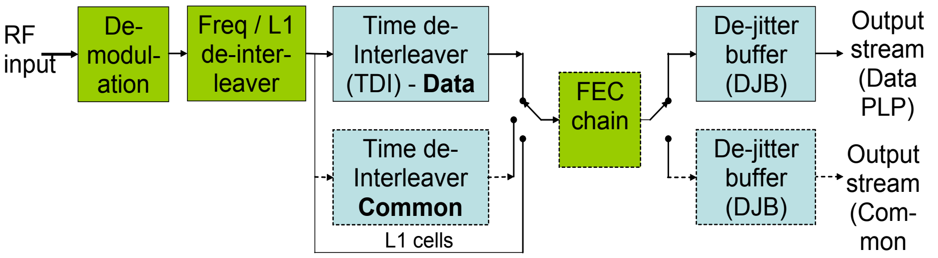


Figure 18: Receiver buffer model

The time de-interleaver receives the output cells from the frequency/L1 de-interleaver stage. For L1 cells no time de-interleaving is applicable. Part of this model is as well a single FEC chain for all three paths to processed, i.e. a single data PLP, the related Common PLP (if any) and the L1 signalling. Outputs of the FEC chain are descrambled BaseBand Frames. The generic form of the data field of the latter shall be the one of normal mode, i.e. 3-byte ISSY and DNP switched on. That data is written into the De-Jitter Buffers (DJBs). Conceptually there are separate memories for the content of the selected Data PLP and the content of the related Common PLP – for time de-interleaving and de-jittering.

The DJB is read out with a rate that is derived from the ISCR values received. This mechanism assures perfect synchronization between modulator sampling rate and read clock. Previously removed sync bytes and deleted null packets are re-inserted as part of this read process. The reading process starts at the point of time that is indicated by the TTO value. The size of the de-jitter buffer for one Data and one Common PLP is 2 Mbit. The size of the whole time de-interleaving memory is $2^{19}+2^{15}$ OFDM cells.

It shall be noted that for particular corner case – e.g. Mode A with ISSY – it helps if there is communication between the TDI and the DJB in order to prevent a DJB overflow and to let the TDI hold the data until the DJB is able to store them.

Null packet deletion

In order not to burden the transmitted multiplex with empty TS packets, those null packets can be removed before transmission, but must be re-inserted by the receiver in order to achieve a constant bit rate and end-to-end delay at the receiver's de-multiplexer. A maximum number of 255 null packets in succession can be removed and the corresponding necessity for re-insertion indicated by means of the related DNP byte transmitted. This DNP parameter signals the number of null packets to be stuffed into the TS again in front of the TS packet it is provided with. This mechanism prevents as well the necessity of a PCR re-stamping.

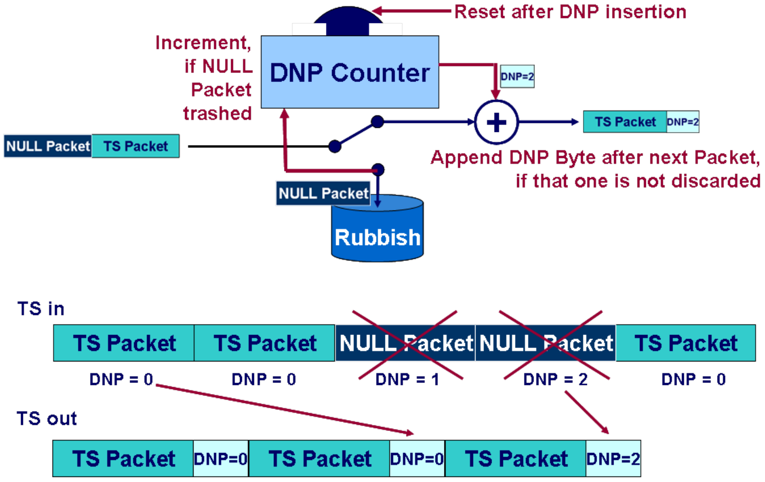


Figure 19: Null packet deletion and re-insertion

4.3.2 Stream adaptation

Regarding the stream adaptation – see Figure 20 below – there is principally no difference between the single and the multiple PLP cases, although there is no real scheduling task in the single PLP case (mode A).

Scheduling

In the multiple PLP case (mode B) the scheduler – together with the Data Slice builder – assigns the cells of each C2 frame to the PLPs to be transported. Also the composition of the Data Slice structure is defined here. Based on those assignments, the scheduler generates the L1-part2 signalling. At this stage the PLPs remain to be unaffected.

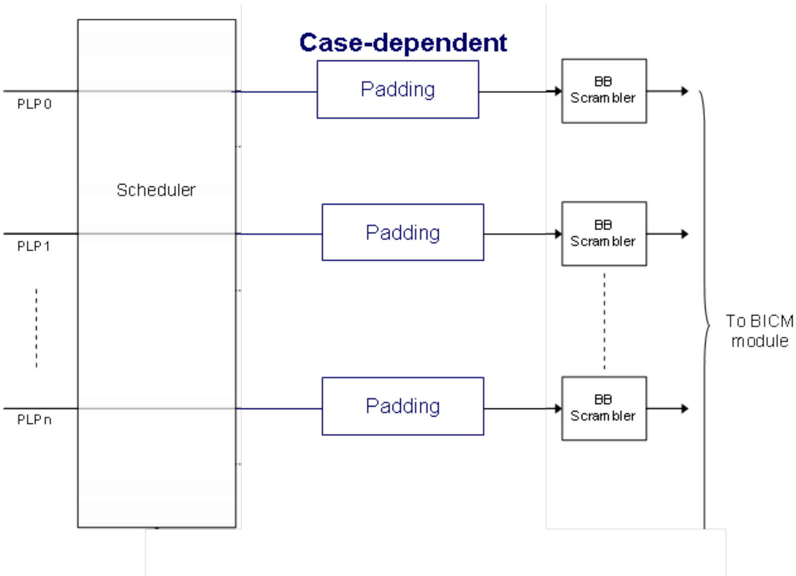


Figure 20: Stream adaptation block

In order to enable L1 signalling in advance of the described data, the latter needs to buffered before transmission. The scheduler counts the number of FECFrames of each PLP at its input and correspondingly determines the dynamic L1 parameter settings and provides those to the L1 signalling generator.

Padding insertion

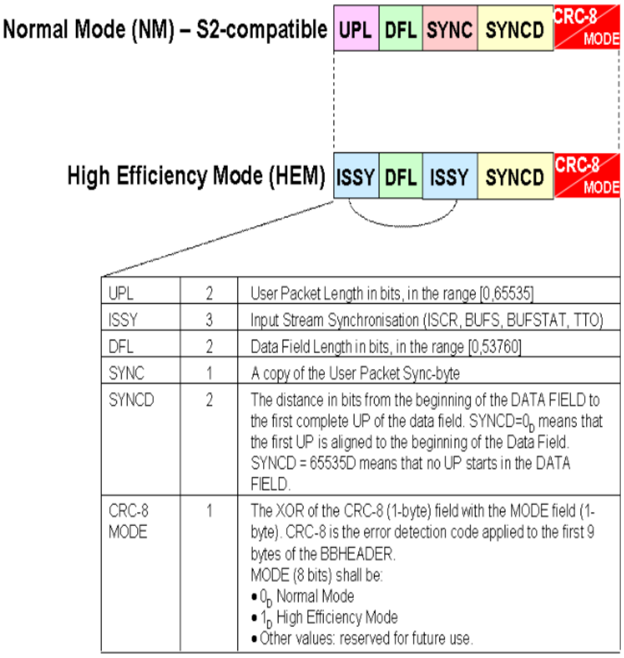
The first functional block inserts padding at the end of the baseband frames where the data field doesn't fill the available space. The length of the data field and the padding field (if present) together depends on the chosen code rate. The padding consists of zero bits only.

Baseband frame scrambling

This stage applies randomization to the content of each baseband frame by 'EXOR'ing this with a PRBS generated with the polynomial $1 + x^{14} + x^{15}$.

4.4 Baseband frame modes

There are two modes available for setting up the stream of BBFs, Normal and High Efficiency Mode (NM, HEM). The related BBF header parts consist of the MATYPE field (3 MSBs, see table 2 below) and the fields outlined with Figure 21 below:



UPL	2	User Packet Length in bits, in the range [0,65535]
ISSY	3	Input Stream Synchronisation (ISCR, BUFS, BUFSTAT, TTO)
DFL	2	Data Field Length in bits, in the range [0,53760]
SYNC	1	A copy of the User Packet Sync-byte
SYNCD	2	The distance in bits from the beginning of the DATA FIELD to the first complete UP of the data field. SYNCD=0_D means that the first UP is aligned to the beginning of the Data Field. SYNCD = 65535D means that no UP starts in the DATA FIELD.
CRC-8 MODE	1	The XOR of the CRC-8 (1-byte) field with the MODE field (1-byte). CRC-8 is the error detection code applied to the first 9 bytes of the BBHEADER. MODE (8 bits) shall be: • 0_D Normal Mode • 1_D High Efficiency Mode • Other values: reserved for future use.

Figure 21: BBF header besides MATYPE (7 LSBs)

MATYPE signalling

The two MSBs of the BaseBand Frame headers (MATYPE field) are used for signalling PLP-specific parameters as well as the presence/absence of optional fields:

TS/GS (2 bits)	SIS/MIS (1 bit)	CCM/ACM (1 bit)	ISSYI (1 bit)	NPD (1 bit)	EXT (2 bits)
00 = GFPS 11 = TS 01 = GCS 10 = GSE	1 = single 0 = multiple	1 = CCM 0 = ACM	1 = active 0 = not active	1 = active 0 = not active	Reserved for future use (see note 1)

NOTE 1: For T2, EXT=reserved for future use and for S2, EXT=RO =transmission roll-off.
NOTE 2: For compatibility with DVB-S2, when GSE is used with normal mode, it shall be treated as a Continuous Stream and indicated by TS/GS = 01.

Table 2: MATYPE-1 field mapping

The principal difference is that for those streams with fixed and a priori known packet length and with known synchronization sequence the Input Stream Synchronization field is moved from the User Packets to the BBF headers replacing there the mentioned fields. This is called the High Efficiency Mode.
The normal mode reflects exactly the corresponding structure of DVB-S2, i.e. is perfectly suitable for direct retransmission in a cable network of the content received via S2.

The following paragraphs illustrate which mode is applicable to which stream type situations and how the BBFs and their content look like for each of the possible cases.

<u>Normal mode</u>
The normal mode looks slightly different for the stream types GFPS/TS on the one hand side and GCS/GSE on the other hand as illustrated with the figures below:

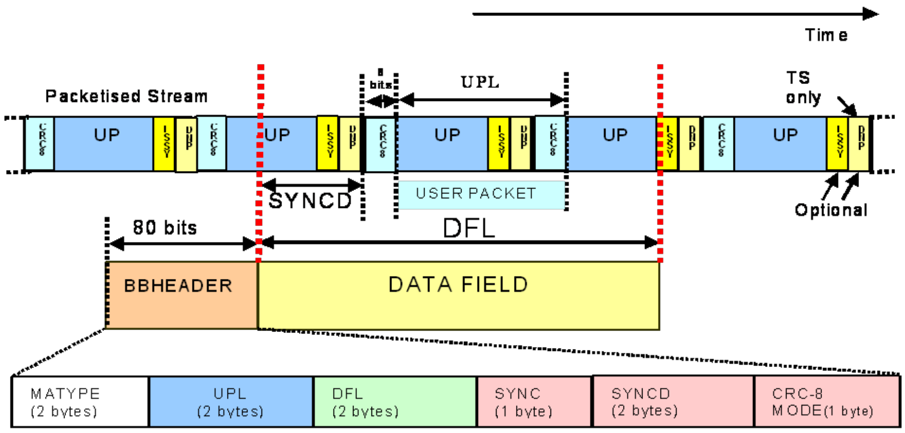


Figure 22: Mode adapter output, Normal Mode, GFPS and TS

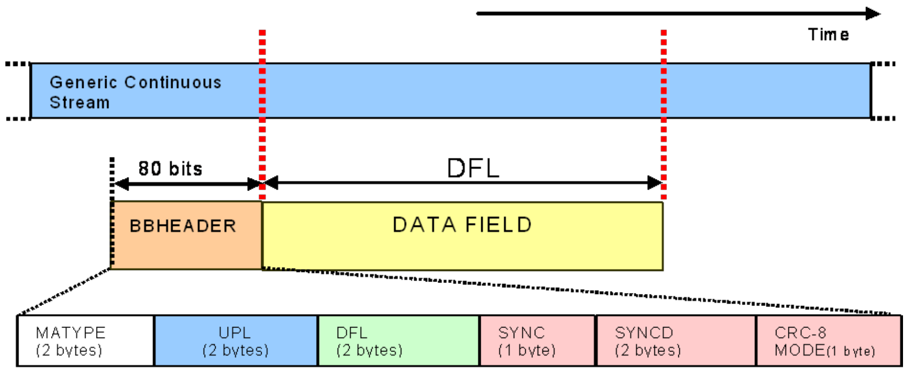


Figure 23: Mode adapter output, Normal Mode, GSE and GCS

For GSE and GCS no additional fields – like ISSY, DNP and CRC-8 – are inserted into the input stream.

<u>High efficiency mode</u>
In a similar way as is the case for the normal mode above, also for the high efficiency mode one needs to distinguish between two stream type case, namely TS and GSE, see the figures below:

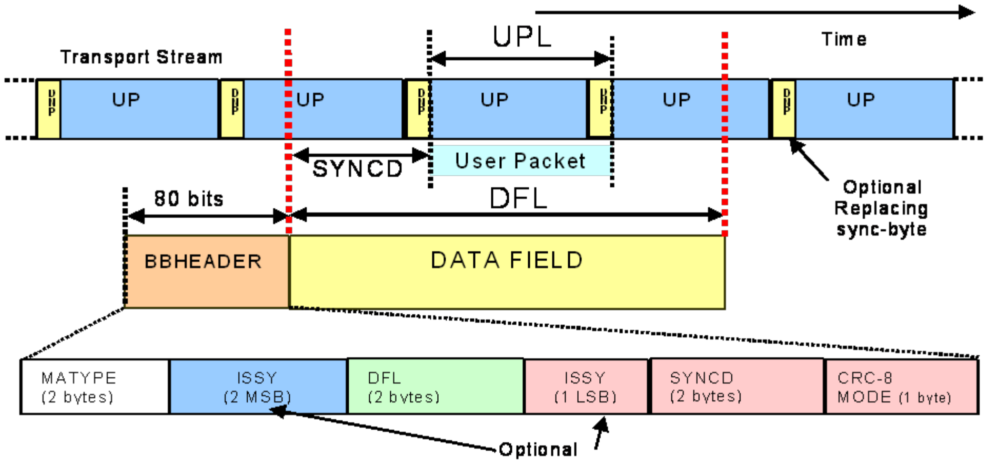


Figure 24: Mode adapter output, High Efficiency Mode for TS (no CRC-8 computed for UPs, optional single ISSY inserted in the Baseband Header, UPL not transmitted)

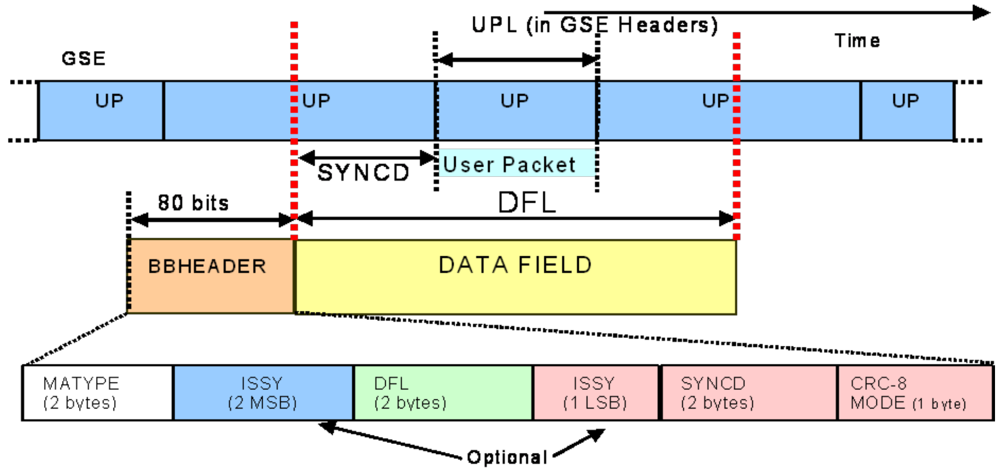


Figure 25: Mode adapter output, High Efficiency Mode for GSE (no CRC-8 computed for UPs, optional single ISSY inserted in the Base-band Header, UPL not transmitted)

Whereas in the TS case the DNP mechanism (see sub-clause 4.3.1.2 for details) is applicable, in the GSE case no such method is applied. Commonalities are the missing CRC-8 field and the substitution of the UPL and SYNC fields by the ISSY parameter.

4.5 Statistical multiplexing for DVB-C2

Traditionally video encoding combined with statistical multiplexing works in a way that the sum of the data rates of the different video streams never exceeds a particular threshold. A typical example would be a DVB-T system with 4 TV services and an upper limit of 12 Mbps for the 4 video streams involved (remaining bit rate of the related multiplex consumed by audio, signalling etc.). Background to this set-up is the fact that broadcasting systems in the past worked with constant coding and modulation. C2 now is able to transport content through individual pipes (PLPs), of which each can be assigned an individual combination of error control coding and constellation /modulation, i.e. variable coding and modulation can employed. Therefore the statistical multiplexer must consider these settings in order to generate slices of video streams that do not exceed what a single C2 frame can carry in terms of cells with individual coding and modulation for each PLP.

It should also be noted that the coding and modulation can be changed from C2 frame to C2 frame and from FECFrame to FECFrame, respectively.

A simple example for the constant cell rate statistical multiplexing is given with Figure 26 below. Here two services are statistically multiplexed. The PLP carrying service 1 (in this simple example consisting only of its video component) is 256-QAM-modulated and enjoys a code rate of ¾, whereas the second PLP carrying service 2 is 64-QAM-modulated and its code rate is ½.

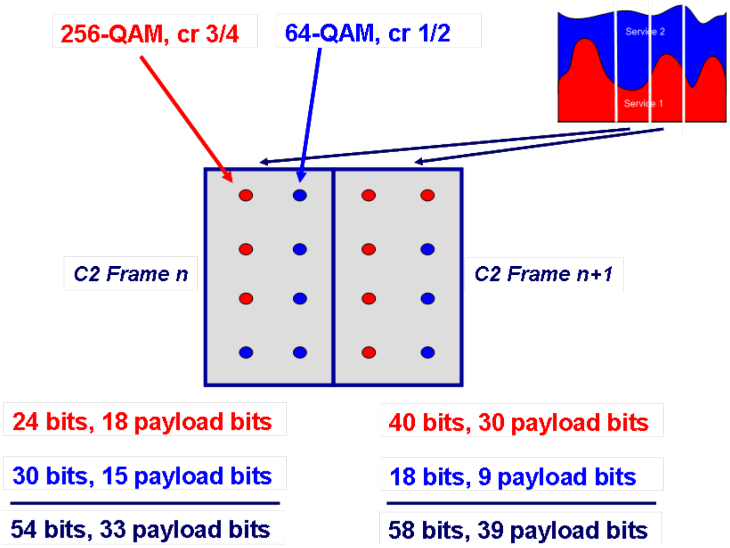


Figure 26: Constant cell rate statistical multiplexing

5 Bit-Interleaved Coded Modulation (BICM)

The term bit-interleaved coded modulation or BICM in short, includes all processing steps involved in encoding a base-band frame into a Data Slice packet. This includes the FEC encoding, various interleaving steps, the QAM modulation and finally the FECFrame header insertion. All the above processing steps are described in this chapter, except for the FECFrame header insertion which is covered in Chapter 6.

DVB-C2 BICM is quite a bit more flexible and powerful compared to DVB-C. The DVB-C BICM utilizes Reed-Solomon FEC encoding at a single code rate and only 3 QAM constellation mappings. The primary advantages of DVB-C2 over DVB-C are:

- DVB-C2 is much closer to the theoretical (Shannon) limit, providing more user data throughput per cable bandwidth
- DVB-C2 provides greater flexibility with a coding & modulation granularity of almost 2 dB
- Higher modulation orders in DVB-C2 enable higher capacity

The previous chapter introduced the concept of multiple independent physical layer pipes (PLPs). In the BICM encoder, each PLP needs to be encoded to individual parameters and as such is seen as a completely autonomous encoding chain.

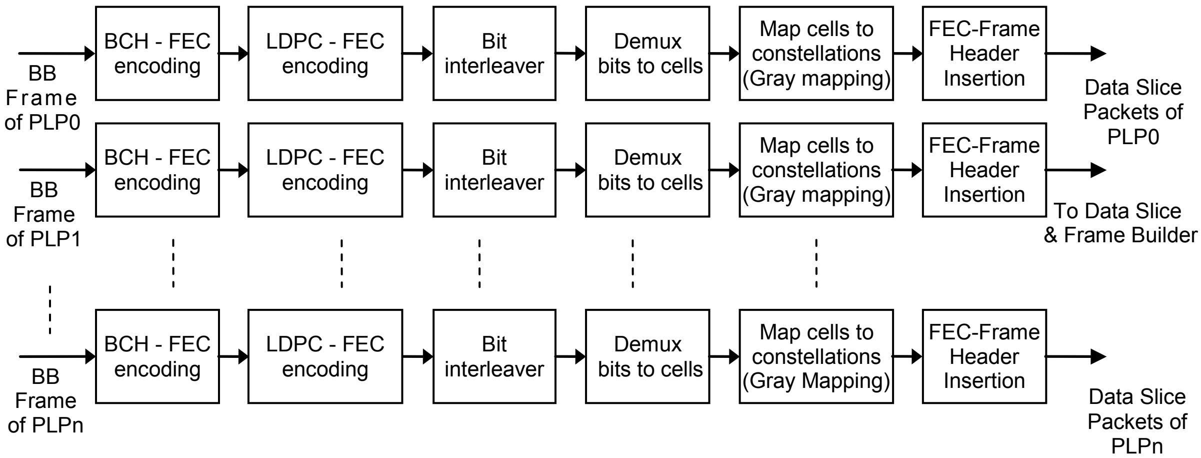


Figure 27: Each PLP has an independent BICM encoder

Figure 27 illustrates the parallel structure of BICM encoding over PLPs. The subsequent elaboration does not take multiple PLPs into account, but please note that all processing needs to be parallelized as needed.

5.1 BICM FEC coding

The combination of a powerful soft decision Low Density Parity Check Code (LDPC) with a low code rate BCH code was first introduced in DVB-S2. The excellent performance which was proven in the field over millions of units made it an obvious choice for DVB-C2. The same LDPC+BCH coding was also adopted for DVB-T2 with the addition one code rate.

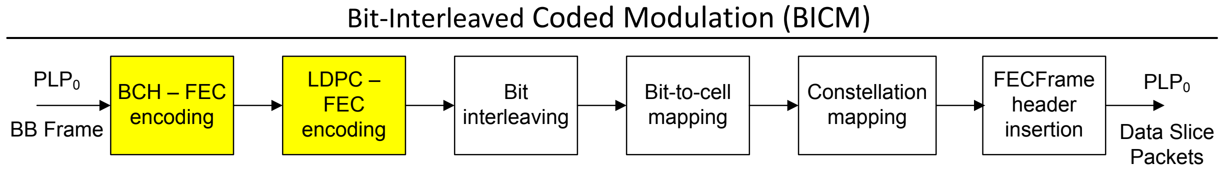


Figure 28: Block diagram of signal processing for the creation of a PLP with the location of the FEC indicated

In a first encoding step, the BCHFEC is coded on the baseband frame (BBFrame) and the LDPCFEC is coded on top of the N_{bch} frame to yield N_{ldpc} bits in the second coding step. As in DVB-S2 and DVB-T2, two FECFrame lengths are specified, the normal FECFrame, with N_{ldpc} = 64800 bits and the short FECFrame with N_{ldpc} = 16200 bits. Figure 29 shows the final FECFrame after coding.

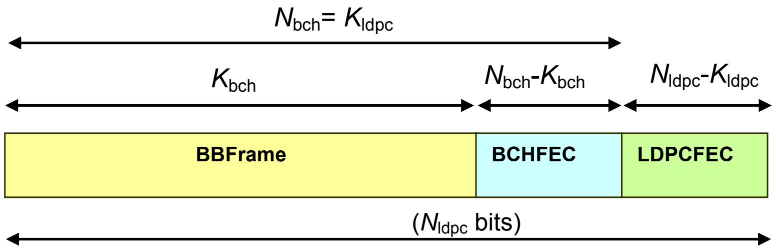


Figure 29: BCH FEC and LDPC FEC coding is added to the BBFrame and forms the FECFrame

5.1.1 Normal FECFrame

The coding options for BCH and LDPC are shown in **Table 3** below. The normal FECFrame length is constant and the uncoded user data varies with BCH and LDPC code rate.

LDPC Code	BCH Uncoded Block K_{bch}	BCH coded block N_{bch} LDPC Uncoded Block K_{ldpc}	BCH t-error correction	N_{bch}-K_{bch}	LDPC Coded Block N_{ldpc}
2/3	43 040	43 200	10	160	64 800
3/4	48 408	48 600	12	192	64 800
4/5	51 648	51 840	12	192	64 800
5/6	53 840	54 000	10	160	64 800
9/10	58 192	58 320	8	128	64 800

Table 3: Coding available for Normal FECFrames

Only a selection of code rates from DVB-S2 is available for the DVB-C2 normal FECFrame coding. The selection was made by plotting the spectral efficiency of all available code rates at all QAM modulation orders against the QEF SNR performance. A commercial requirement is to have a granularity of less than 3dB in the code rate / modulation range. The second selection criterion was to give preference to lower QAM constellation orders. For example, 1024-QAM code rate 9/10 and 4096-QAM code rate 3/4 yield similar spectral efficiency / SNR performance, so the 1024-QAM modulation was chosen. Figure 30 illustrates the code rate choices available in DVB-C2 at the various QAM constellations. It can be clearly seen that all DVB-C2 codes are much closer to the Shannon limit than the DVB-C codes.

The operator has much more flexibility to adjust the coding to the particular cable performance values. When replacing DVB-C with DVB-C2 transmission, it is possible to either improve the throughput, increase the robustness or a trade-off of the two improvements.

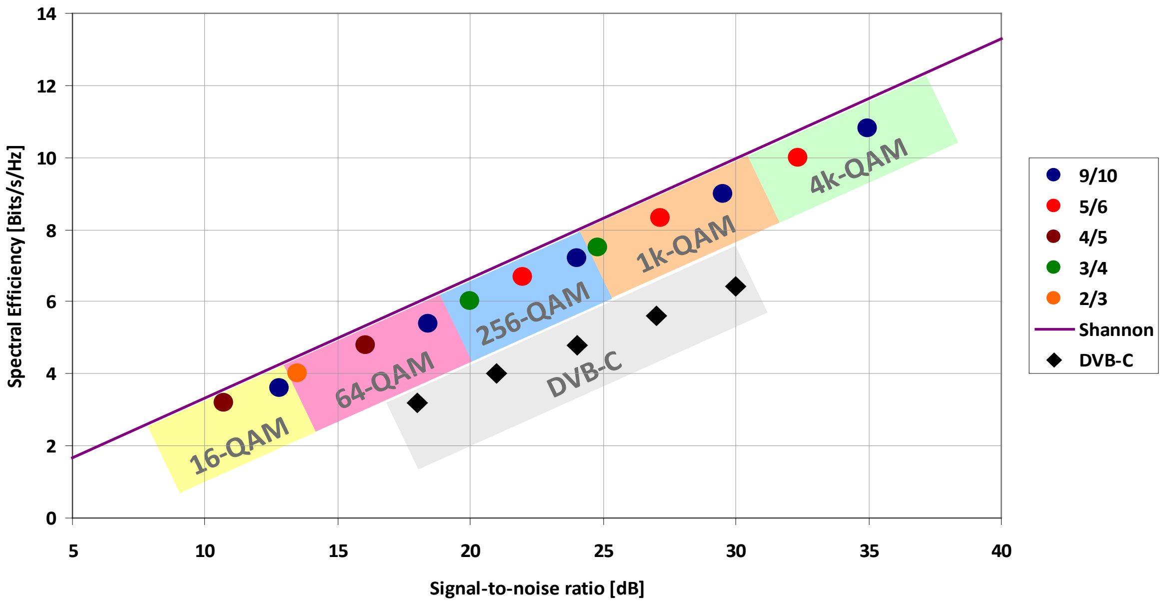


Figure 30: Normal FECFrame coding granularity

For example, DVB-C 256-QAM needs to have at least 30dB SNR for QEF transmission. It is possible to increase the throughput by ~40% by moving to DVB-C2 1024-QAM code rate 9/10 at the same SNR requirements of ~30dB. Alternatively, the throughput can be kept constant and the robustness improved to only a 20dB SNR requirement by moving to DVB-C2 256-QAM code rate 3/4.

The 13 DVB-C2 Normal FECFrame code rate versus QAM-constellation coding options are summarized in Table 4 and compared to the 3 DVB-C modes.

Mode	**16-QAM**	**64-QAM**	**256-QAM**	**1024-QAM**	**4096-QAM**
DVB-C	25.6 Mbps	38.4 Mbps	51.2 Mbps	×	×
DVB-C2: 2/3	×	31.4 Mbps	×	×	×
DVB-C2: 3/4	×	×	47.1 Mbps	58.9 Mbps	×
DVB-C2: 4/5	25.1 Mbps	37.7 Mbps	×	×	×
DVB-C2: 5/6	×	×	52.4 Mbps	65.4 Mbps	78.6 Mbps
DVB-C2: 9/10	28.3 Mbps	41.4 Mbps	56.6 Mbps	70.7 Mbps	84.8 Mbps

Table 4: Throughput of all DVB-C2 normal FECFrame modes for 8 MHz transmission

5.1.2 Short FECFrame

The short FECFrame of 16200 bits is used both for payload data and to protect l1 pre-signalling. At any given code rate, short FECFrame codes are less robust than the normal FECFrame codes because the code length has a direct influence in the LDPC performance. Through the smaller packet size makes short FECFrames well suited for applications where a low latency is required.

LDPC Code Identifier	BCH Uncoded Block K_{bch}	BCH Coded Block N_{bch} LDPC Uncoded Block K_{ldpc}	BCH t-error correction	N_{bch}-K_{bch}	Effective LDPC Rate $K_{ldpc}/16,200$	LDPC Coded Block N_{ldpc}
1/2 *	7 032	7 200	12	168	4/9	16 200
2/3	10 632	10 800	12	168	2/3	16 200
3/4	11 712	11 880	12	168	11/15	16 200
4/5	12 432	12 600	12	168	7/9	16 200
5/6	13 152	13 320	12	168	37/45	16 200
8/9	14 232	14 400	12	168	8/9	16 200

* **NOTE:** This code rate is only used for protection of L1 pre-signalling and not for data

Table 5: Coding available for Short FECFrames

The coding options for BCH and LDPC are shown in Table 5. The short FECFrame length is constant and the uncoded user block length varies with BCH and LDPC code rate.

The available DVB-C2 short FECFrame modes are shown in Table 6 below.

	Short FECFrame Modulation formats					
Code rate	**QPSK**	**16-QAM**	**64-QAM**	**256-QAM**	**1024Q-AM**	**4096-QAM**
1/2	✓	✓	-	-	-	-
2/3	✓	-	✓	-	-	-
3/4	✓	-	-	✓	✓	-
4/5	✓	✓	✓	-	-	-
5/6	✓	-	-	✓	✓	✓
8/9	✓	✓	✓	✓	✓	✓

Table 6: Code rate – Constellation selection for short FECFrames

5.2 BICM bit interleaving

The bit interleaving follows the FEC processing as indicated in Figure 31.

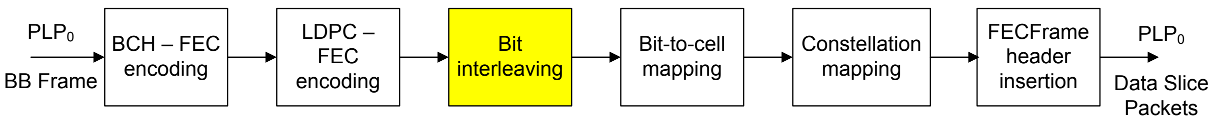


Figure 31: Block diagram of signal processing for the creation of a PLP with the location of the bit interleaving indicated

Each FECFrame is parity interleaved and then bit-interleaved using a column twist algorithm. Parity interleaving is taken from DVB-S2 and DVB-T2 without modification.

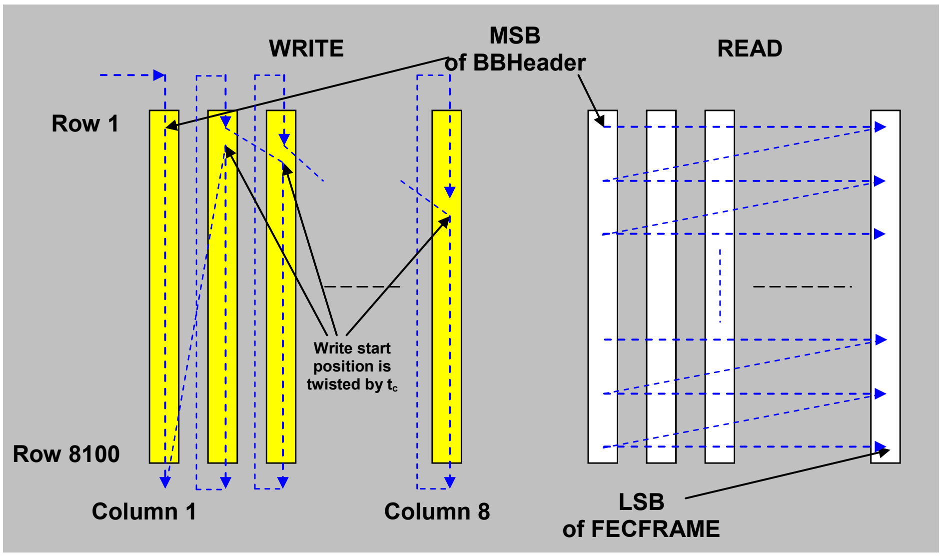


Figure 32: Column-twist bit interleaving

The column-twist bit-interleaver settings for 16-, 64- and 256-QAM established in the DVB-T2 standard were re-used for DVB-C2. New DVB-C2 parameters were defined for 1024-QAM and 4096-QAM. Table 7 below lists the geometry of all bit-interleavers by constellation and FECFrame length.

Modulation	**Rows N_r**		**Columns N_c**
	N_{ldpc} **= 64, 800**	N_{ldpc} **= 16,200**	
16-QAM	8 100	2 025	8
64-QAM	5 400	1 350	12
256-QAM	4 050	-	16
	-	2 025	8
1024-QAM	3 240	810	20
4096-QAM	5 400	-	12

Table 7: Column twist geometries

Bit interleaving together with bit-to-cell mapping have a large impact on the error floor performance. The bit interleaver configurations incorporated in DVB-C2 were thoroughly simulated and tested to optimize the error floor.

Table 8 lists the new DVB-C2 column-twist parameters for 1024- & 4096-QAM for both normal and short FECFrames. All parameters were optimized to minimize the error-floor level.

	Normal FECFrame		Short FECFrame	
	1024-QAM	**4096-QAM**	**1024-QAM**	**4096-QAM**
N_c	**20**	**12**	**20**	**24**
0	0	0	0	0
1	1	0	0	0
2	3	2	0	0
3	4	2	2	0
4	5	3	2	0
5	6	4	2	0
6	6	4	2	0
7	9	5	2	1
8	13	5	5	1
9	14	7	5	1
10	14	8	5	2
11	16	9	5	2
12	21	-	5	2
13	21	-	7	3
14	23	-	7	7
15	25	-	7	9
16	25	-	7	9
17	26	-	8	9
18	28	-	8	10
19	30	-	10	10
20	-	-	-	10
21	-	-	-	10
22	-	-	-	10
23	-	-	-	11

Table 8: New column-twist parameters for 1024-QAM & 4096-QAM

5.3 BICM Bit-to-Cell Mapping

After the bit interleaving unit described above, the bit-to-cell mapping is implemented (see Figure 33).

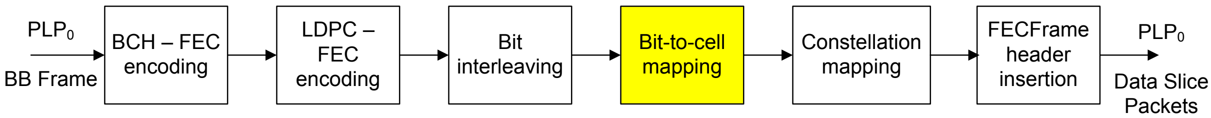


Figure 33: Block diagram of signal processing for the creation of a PLP with the location of the bit-to-cell mapping indicated

The bit-to-cell mapping maps the continuous FECFrame bit-stream into individual QAM cells. This increases the robustness of the higher order QAM modulations as the bits which are weakly protected are scrambled on the constellation plot. Figure 34 shows that such a mapping can be implemented by means of a de-multiplexing processing.

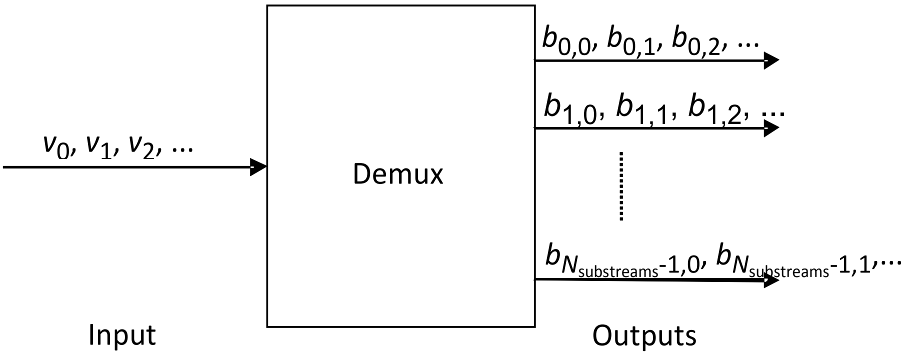


Figure 34: Implementation of bit-to-cell mapping by means of a de-multiplexer

The bit-to-cell mapping structure for 16-, 64-, and 256-QAM were directly re-used from the DVB-T2 standard, except for 256-QAM Normal frame using code rate 2/3. New mappings for 1024-QAM and 4096-QAM for both normal and short FECFrames are presented in DVB-C2.

5.3.1 Normal FECFrame

Figure 35 below shows how each bit of the 20 bits from a normal FECFrame are mapped to two 10-bit wide 1024-QAM symbols.

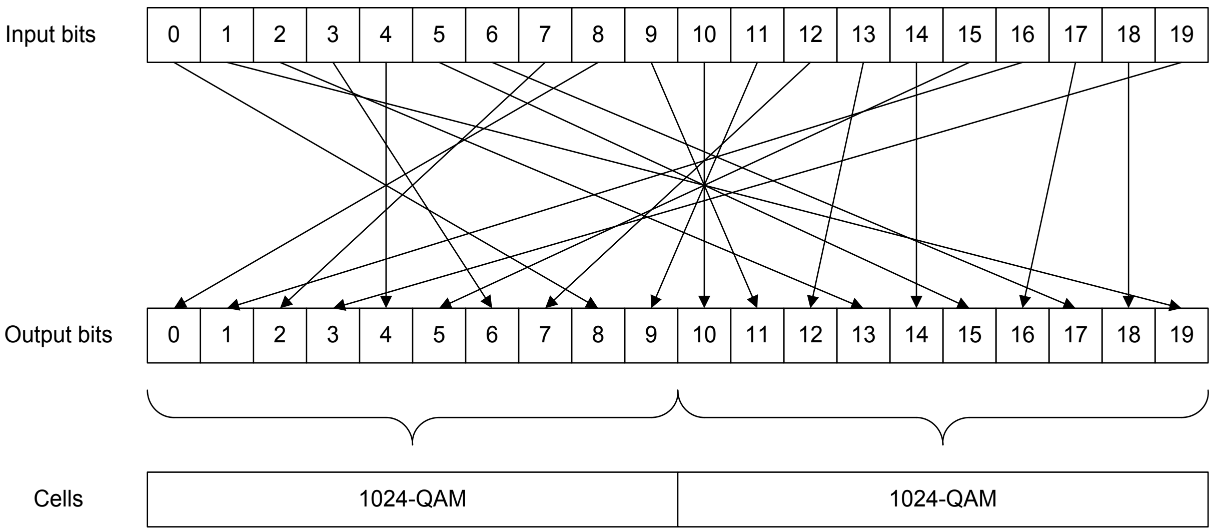


Figure 35: 1024-QAM normal FECFrame bit-to-cell mapping.

In 4096-QAM mode, a normal FECFrame mapping is applied by which every 12 bits are mapped to a single 4096-QAM symbol. The mapping rule is indicated in Figure 36.

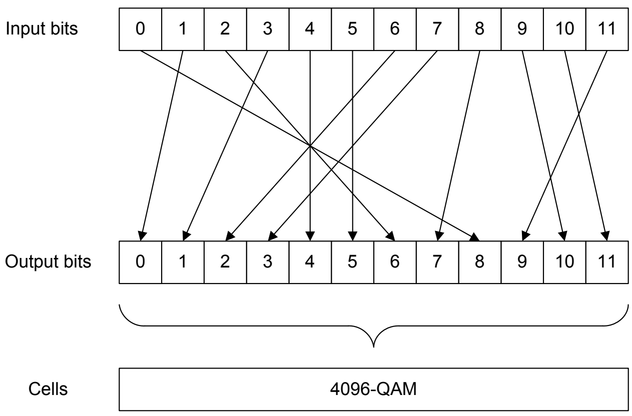


Figure 36: 4096-QAM normal FECFrame bit-to-cell mapping.

5.3.2 Short FECFrame

The first DVB-C2 standardization approach was to use the normal FECFrame bit-to-cell mapping also for the short FECFrames. However, hardware simulations showed that this yielded poor error floor performance and the following optimized bit-to-cell mappings were introduced for the short FECFrame lengths. Figure 37 shows the 1024-QAM short FECFrame mapping where 20 input bits are mapped to 2 consecutive 1k-QAM cells.

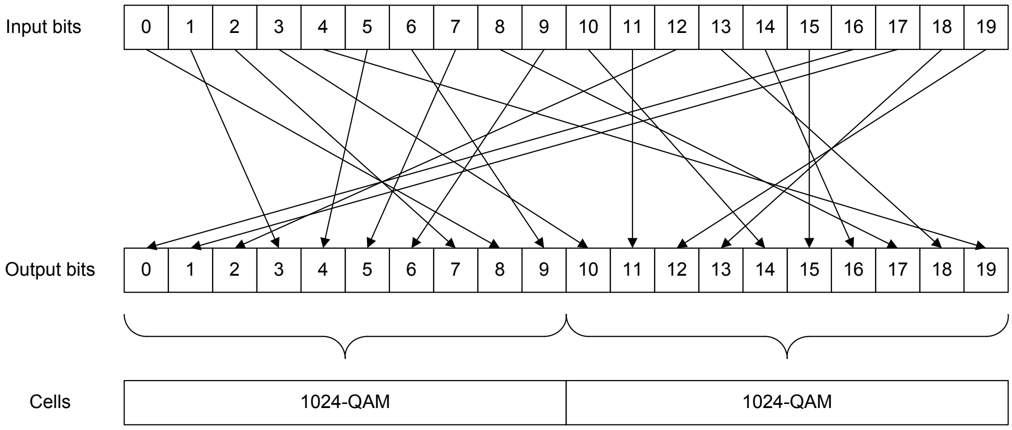


Figure 37: 1024-QAM short FECFrame bit-to-cell mapping.

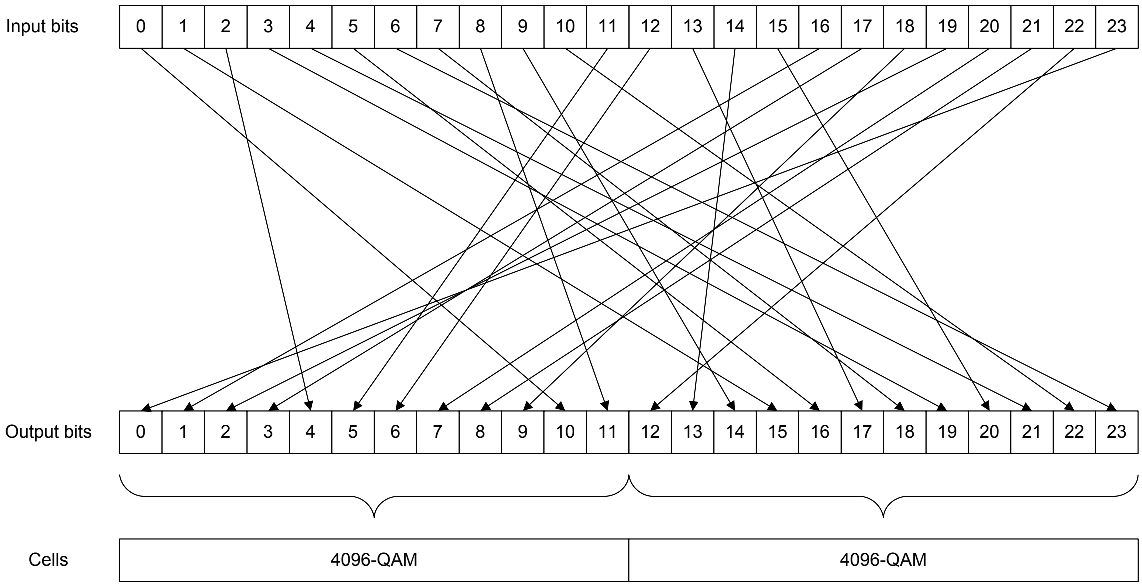


Figure 38: 4096-QAM short FECFrame bit-to-cell mapping.

Figure 38 shows the 4096-QAM short FECFrame mapping where 24 input bits are mapped to two consecutive 4k-QAM cells.

5.4 BICM constellation mapping

The Constellation mapping follows the bit-to-cell mapping as indicated in Figure 39.

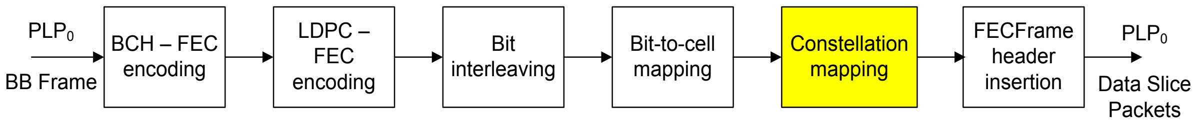


Figure 39: Block diagram of signal processing for the creation of a PLP with the location of the constellation mapping indicated

In the final BICM step, the cells are mapped to constellation points using standard grey mapping. Figure 40 shows the example of a 256-QAM constellation.

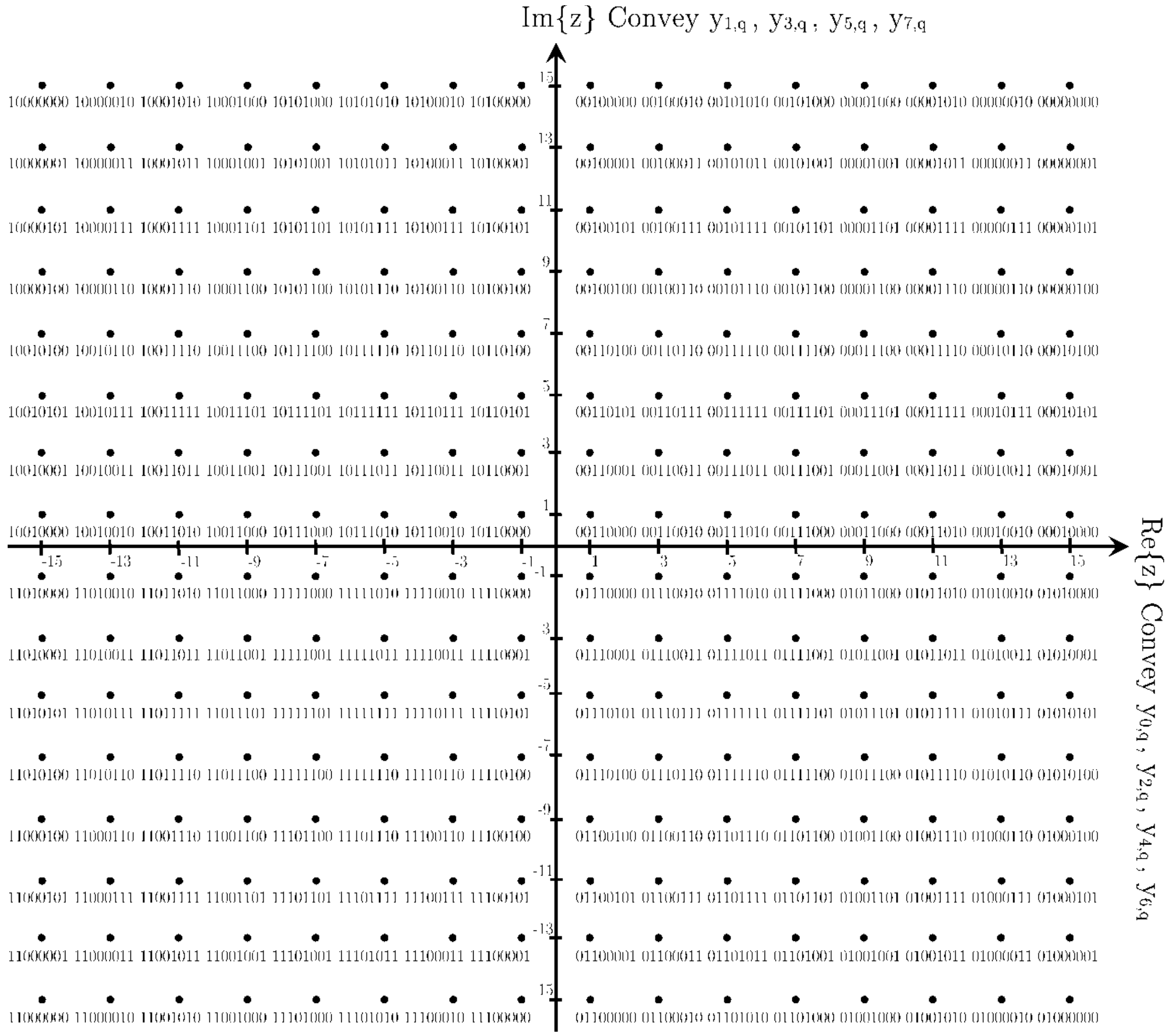


Figure 40: Example for grey mapping of a 256-QAM constellation

The insertion of the FECFrame header is typically included in the term BICM and will be covered in Chapter 6.

6 PLPs and Data Slices

6.1 Location within the DVB-C2 System

The DVB-C2 system is able to transmit multiple independent data streams (e.g. MPEG-2 Transport Streams [1] or Generic Streams [2]) in so-called Physical Layer Pipes (PLP), which can be modulated and error-protected independently from each other. One or multiple PLPs are then multiplexed into so-called Data Slices. These Data Slices have certain similarities to DVB-C channels, but without any frequency spacing between two of them. Hence, the interleaving is performed within Data Slices only, and the reception of a single Data Slice is sufficient for receiving a service (exceptions are the Common PLP or PLP Bundling for higher data rates). Afterwards, one ore multiple Data Slices, together with the Layer 1 signalling data and additional reference information, form the output signal. The complete transmitter block diagram and the locations of the Data Slices and PLPs are depicted in Figure 41.

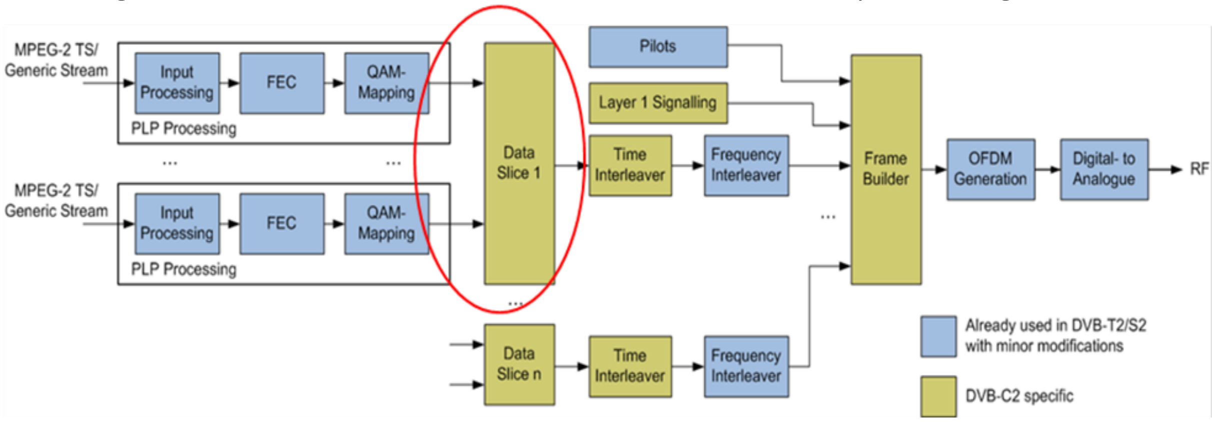


Figure 41: Simplified DVB-C2 transmitter block diagram

6.2 Background behind the PLP and Data Slices Concept

The main goal of the DVB-C2 development was an increase of the spectral efficiency. On the one hand this goal is reached by means of higher QAM constellations and the increased performance of the new powerful forward error correction (FEC). However, on the other hand reduced overhead also increases the available payload bit rate. One means is the application of OFDM (Orthogonal Frequency Division Multiplex), which is already known from terrestrial transmissions standards like DAB or DVB-T. Figure 42 shows the reason for this increased spectral efficiency. Single-carrier modulation, as e.g. applied by DVB-C, requires spectral shaping filters limiting the bandwidth of the signal and performing spectral shaping. In case of DVB-C a root-raised-cosine filter employing a roll-off factor of 0.15 (that causes the same amount in overhead, i.e. 15%) is used. This factor is practically independent of the actual signal bandwidth and cannot be reduced significantly below 10%, as this would e.g. increase the required timing accuracy of the receiver. In contrast, OFDM does not require such a roll-off filter, because the spectral shaping is achieved by the steep shoulders at the frequency edges of OFDM itself. However, even though the attenuation at the edges is quite high, additional Guard Bands are required for separation to neighbouring channels. Additionally, the usage of a Guard Interval and frequency domain pilots for channel estimation are normally required for OFDM. However, due the short echo length within the cable networks the overhead due to pilots and the Guard Interval is very limited (e.g. 2%).

Furthermore, the Guard Bands of approx. 200 kHz are only required at the edges of the spectrum. Thus, the relative spectral loss due to the Guard Bands reduces for broad signals, which is one of the key advantages of OFDM in cable networks. This results in a spectral loss of e.g. 2% for 32 MHz wide channels for DVB-C2, instead of 15% for DVB-C.

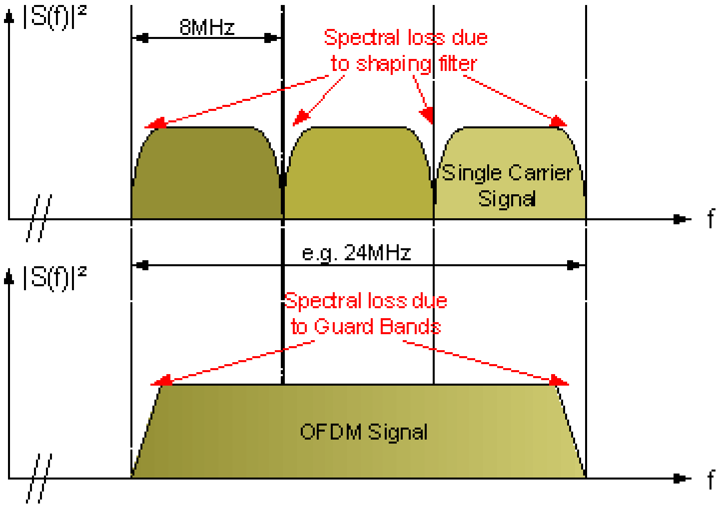


Figure 42: Increased spectral efficiency by means of broadband OFDM

However, due to commercial requirements the possibility to receive the signals using a standard 8 MHz frontend had to be kept, which is reached by means of the Data Slices. Due to the narrow bandwidth of each individual OFDM subcarrier, the receiver is able to demodulate only spectral parts of the OFDM signal, i.e. the Data Slices. As the components of one service are always transmitted within a single Data Slice, the reception of one Data Slice is sufficient to display the service. Therefore, the bandwidth of the Data Slices does not exceed 3408 OFDM subcarriers or 7.61 MHz (for European networks with 8 MHz channel raster) or 5.71 MHz (for US cable networks with 6 MHz channel raster), which allows for the application of receiver front-ends supporting traditional bandwidth (see Figure 43).

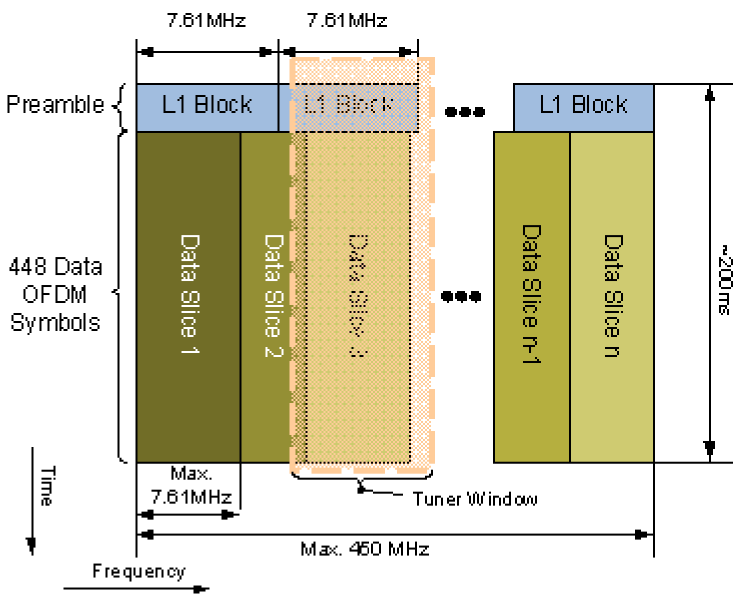


Figure 43: Framing structure of one DVB-C2 Frame in the time and in the frequency domain

One additional requirement is the possibility to have different levels of robustness, which is obtained by means of the Physical Layer Pipes (PLP), as depicted in Figure 44. Each PLP can have its individual modulation and coding parameters that range from Signal-to-Noise Ratios (SNR) of 10 dB to approx. 36 dB. Naturally, this concept makes limited sense for broadcasting services as all cable subscribers shall be able to receive all services. However, it allows for highest spectral efficiency in case of point-to-point services, which are getting more and more important in state-of-the-art cable networks. Examples for these services are Internet access or Video on Demand. As a return channel from the user back to the cable head-end is required anyway, the receiver can also inform the cable head-end about the actual signal quality in the downlink. Hence, the head-end is able to choose the highest possible data rate for each user independently, and the cable spectrum is used most efficiently.

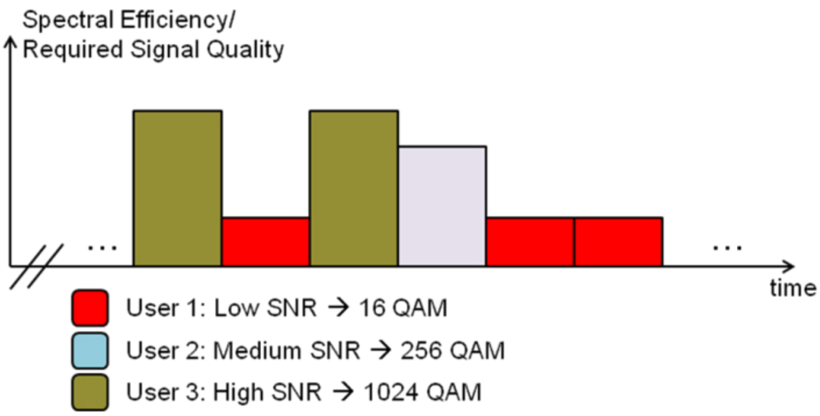


Figure 44: Concept of Physical Layer Pipes with individual levels of robustness

6.3 Types of Data Slices and multiplexing of PLPs

DVB-C2 allows for two types of Data Slices, i.e. Type 1 that is especially suited for broadcast data and Type 2 that is especially suited for interactive services. The Data Slice Type 1 has a fixed parameter set within a frame and allows for the transmission of a single PLP per Data Slice only, but at very low overhead. In contrast, the Data Slice Type 2 allows for the transmission of multiple PLPs per Data Slice with varying modulation and coding parameters, even within a single PLP, at the cost of slightly higher signalling overhead.

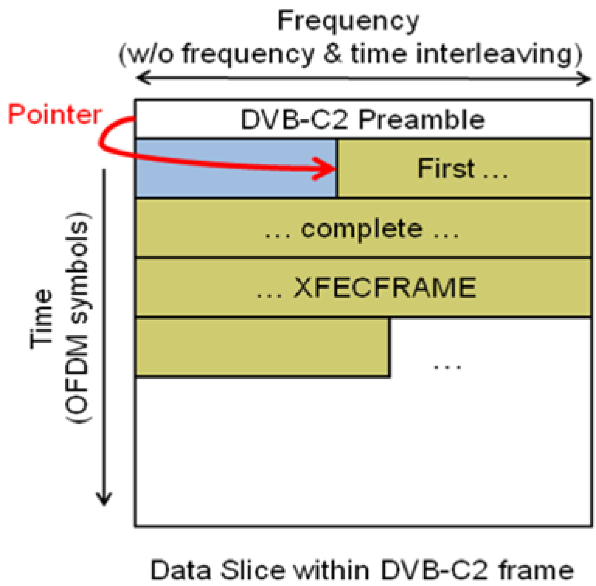


Figure 45: Mapping of PLPs onto Data Slices Type 1 (before frequency interleaving, without pilots)

6.3.1 Data Slice Type 1

Figure 45 depicts the mapping of XFECFRAMEs (i.e. QAM modulated FEC blocks) into the Data Slice Type 1. The data is multiplexed into the Data Slice XFECFRAME by XFECFRAME.

The signalling of the employed modulation and coding parameters, as well as the PLP number, is done within the DVB-C2 Preamble at the beginning of each frame. A pointer field in this preamble points to the first complete XFECFRAME starting within the frame. Additionally, the signalling of the parameters gets valid for the first complete XFECFRAME and is static within this frame. Hence, the fraction of an XFECFRAME already started in the previous C2 frame still uses the parameters of the previous C2 frame.

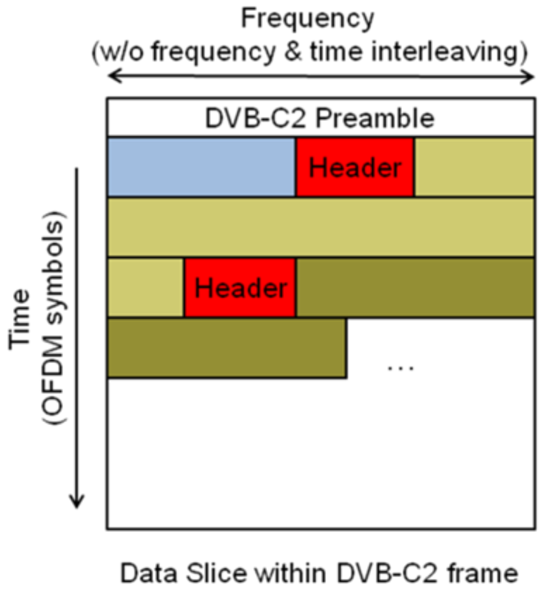


Figure 46: Mapping of PLPs onto Data Slices Type 2

PLP_FEC_TYPE = 0	
PLP_MOD	**XFECFRAME_LENGTH**
000	NA
001	4050
010	2700
011	2025
100	1620
101	1350
110	1158
111	1013

PLP_FEC_TYPE = 1	
PLP_MOD	**XFECFRAME_LENGTH**
000	900
001	16200
010	10800
011	8100
100	6480
101	5400
110	4629
111	4050

Table 9: Length of XFECFRAME depending on the PLP_FEC_TYPE (16K or 64K) and the PLP modulation parameters (e.g. 16-QAM, 64-QAM etc.)

The starting positions of the following XFECFRAMEs can be calculated by means of the modulation and coding parameters in addition to Table 9, as parameters do not change within a frame.

6.3.2 Data Slice Type 2

The Data Slice Type 2 is mainly intended for interactive services. The main reason for this is the reduced latency by means of the signalling directly in front of each XFECFRAME and the increased flexibility. Furthermore, the DVB-C2 Preamble can be completely static, which makes it an interesting alternative for small cable head-ends.

6.3.2.1 *Structure of Data Slice Packets for Data Slices Type 2*

The Data Slice Packets for the Data Slice Type 2 consist of a FECFRAME header and one or two XFECFRAMEs carrying the payload data. The parameters of the following XFECFRAME(s) are signalled within the FECFRAME header. It consists of 32 QPSK symbols (robust mode) or 16 16-QAM symbols (high efficiency mode), which transmit the 16 signalling bits. The application of the QPSK or the 16-QAM header is signalled within the DVB-C2 preamble and is constant for each Data Slice.

The FECFRAME header signals the parameters PLP ID, code-rate, QAM scheme, and if one or two XFECFRAMEs (with the same parameters PLP ID, code-rate and QAM scheme) follow the FECFRAME header. Especially the transmission of the PLP ID allows the receiver to detect if the following XFECFRAME is required to demodulate the desired service. Other XFECFRAMEs do not have to be demodulated, which can save power in the receiver chain.

The transmission of the data within the FECFRAME header has to be as robust as possible, because a disturbed header will lead to complete loss of the payload data. Therefore, DVB-C2 offers two modes, the robust and the high efficiency header transmission. If the more robust payload modes (e.g. 16-QAM or 64-QAM) are employed, the robust header shall be used.

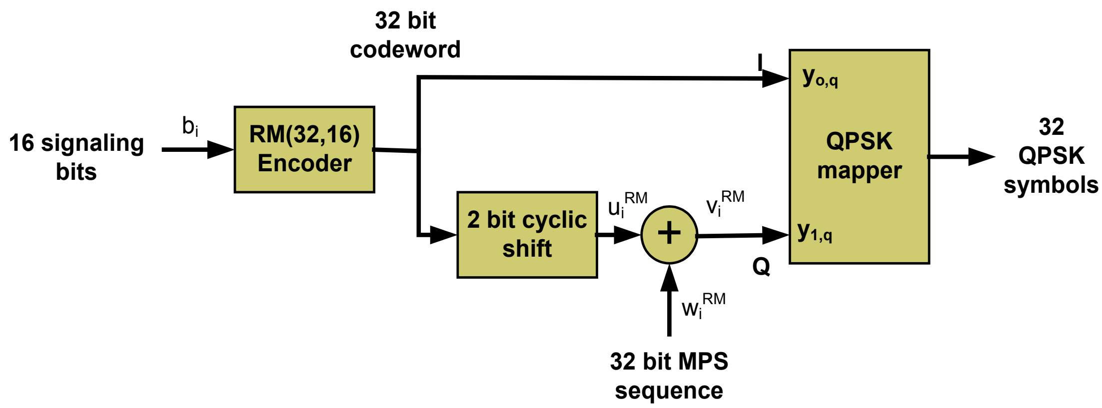


Figure 47: Encoding of the 16 signalling bits for the robust (QPSK) header mode

Figure 47 depicts the encoding of the robust preamble (the high efficiency preamble is encoded in a similar way). The 16 signalling bits are encoded using a code-rate 1/2 Reed Muller code. In case of the QPSK header all 32 coded bits are transmitted on the in-phase (I) component of the 32-QPSK symbols. Additionally, the same 32 bits are transmitted on the quadrature component (Q). However, these bits are cyclically shifted by two bits within the 32 bits (i.e. bit 1→3, bit 2→4, …, bit 31→1, bit 32→2). Furthermore, these bits are scrambled using a fixed 32 bit long MPS sequence.

6.3.2.2 *Synchronisation to Data Slice Packets for Data Slice Type 2*

The DVB-C2 preamble does not signal the start of the Data Slice Packets for the Data Slice Type 2. Hence, the preamble has to provide the synchronisation capabilities on its own. This aim is reached by means of the transmission of the signalling data on the I- and Q-axes of the QPSK diagram (similar for 16-QAM) and the scrambling due to the PRBS sequence.

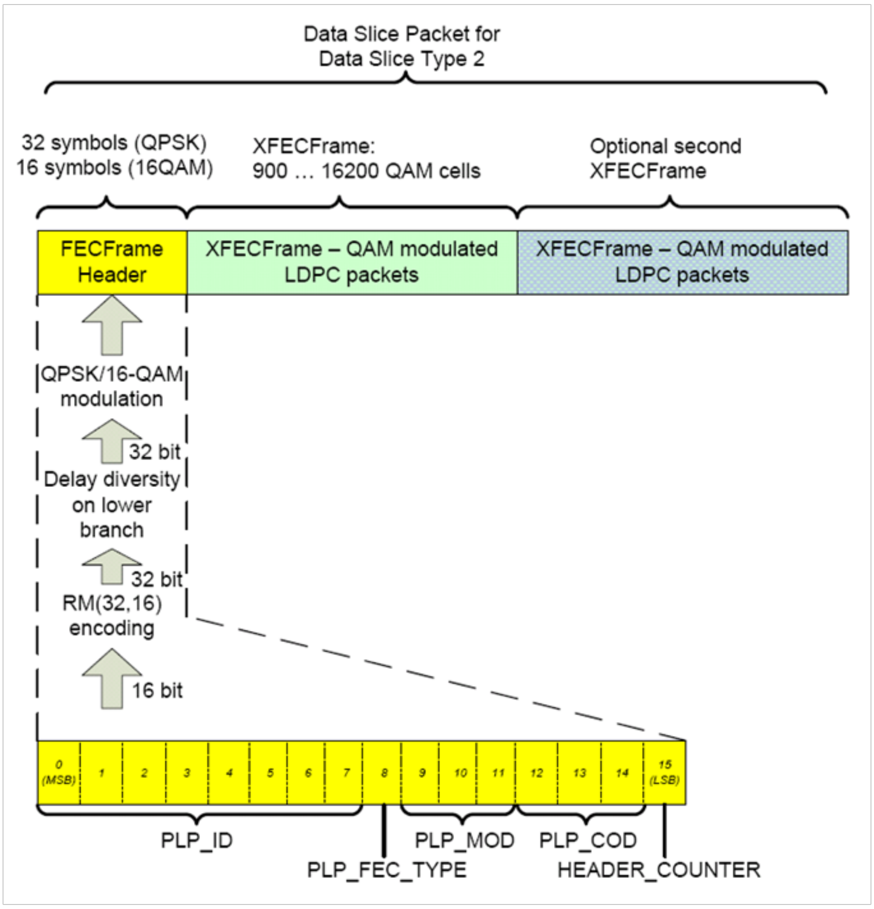


Figure 48: Structure of Data Slice Packets for Data Slice Type 2

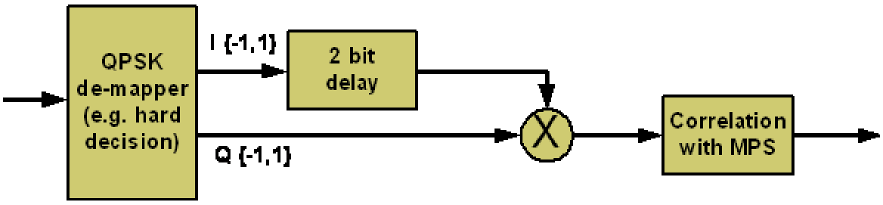


Figure 49 shows the synchronisation to the headers for the robust mode. As the equalisation of the channel has already been done in previous stages of the receiver, it is able to de-map the QPSK symbols. In contrast to the encoding of the header, the delay is now included in the in-phase (I) branch. When the de-mapped data of both branches is multiplied by each other, the output is exactly the MPS sequence for the error-free case. Naturally, the complete sequence is not obtained, as the delay does not completely remove the cyclic shift. However, the length should be sufficient to correlate the signal against the MPS sequence. The detected peak then allows estimating the beginning of a Data Slice Packet correctly.

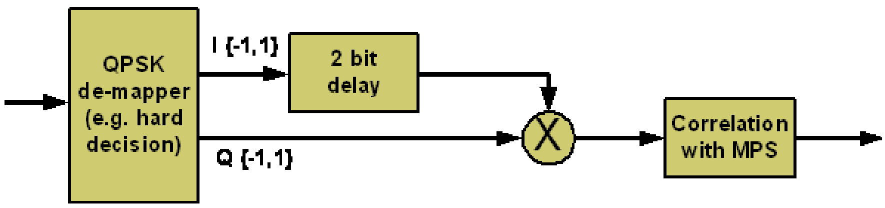


Figure 49: Synchronisation scheme to robust header

After the synchronisation to the header is reached, the signalling data has to be decoded. One approach is shown in Figure 50. In order to achieve the highest performance, the cyclic shift and the MPS sequence has to be removed. For highest performance the complete processing should be done using soft bits (e.g. Log Likelihood Ratio values). This allows for combining the two branches in an optimal way to get optimum diversity. Afterwards, the Reed-Muller decoder removes the remaining errors.

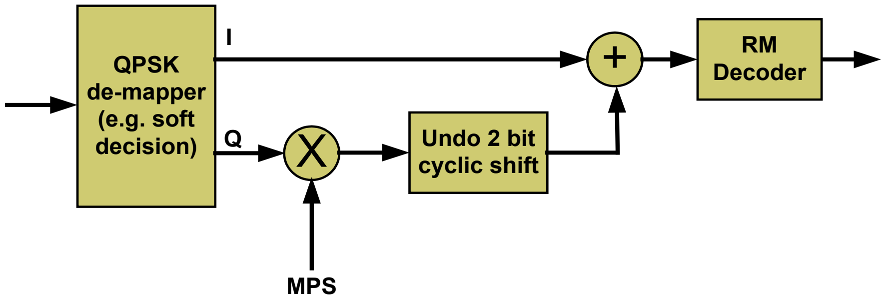


Figure 50: Decoding of the signalling data

Whilst the decoding of the decoding of the Data Slice Packet Headers is required for every Data Slice Packet, the synchronisation to the header is only required once. Afterwards, the position of the following header can be obtained by means of the signalling data within the header in addition to, as this allows for the calculation of the actual XFECFRAME length.

Figure 51: Calculation of the position of the position of the following Data Slice Packet Header by means of the header data

7 Time Interleaving

The background of using Time Interleaver (TI) for DVB-C2 system can be well understood by investigating the characteristics of the cable channels. The important model of cable channel is first provided by IEEE 802.14, which shows the burst noise characteristics of the cable transmission like:

(1) One event per minute with a duration of up to 30 µs
(2) Peak level of +20 dB above signal level

A more recent research result reported by ReDeSign reveals the cable environment becomes better than the past [3]. However, the channel status depends on the country to deploy DVB-C2 system as its digital cable transmission system. So the functionality to overcome such a vulnerable point seems crucial for successful deployment of the DVB-C2 system

Another important point to be addressed is the quite large size of the constellation used by the C2 system. It uses even 4096-QAM constellations and makes it possible to carry several FEC block by a single OFDM symbol. As a result, if the burst noise occurs during the transmission thus degrades the quality of the relevant OFDM symbol, the damage is imposed to the whole FEC blocks rather than the part of them. This may cause unwanted error floor even after channel decoding since those damaged FEC blocks might have no reliable information enough to correct the errors. All the reasoning above is one of the concerns of the DVB TM-T2 group and push them to put the TI functionality into C2 specification.

The architecture relevant to TI function in the Data Slice and L1 signalling part2 is depicted in Figure 52. The TI always precedes the frequency interleaver, which guarantees any FEC blocks output from the Data Slice builder or L1 signalling part2 encoder are optimally spread into whole range of time and frequency region.

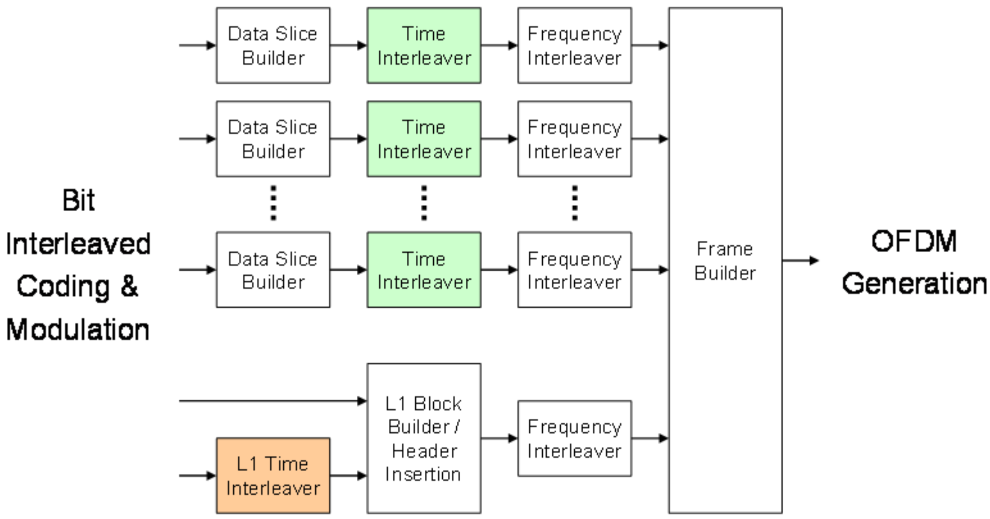


Figure 52: System architecture relevant to time interleaver

Figure 53 shows the typical example of the Data Slice structure which uses the TI function. There are two Data Slices having different width separated by the dotted blue lines. In the (m-1) Data Slice, there are several PLPs and a specific PLP among them is indicated by orange FEC blocks.

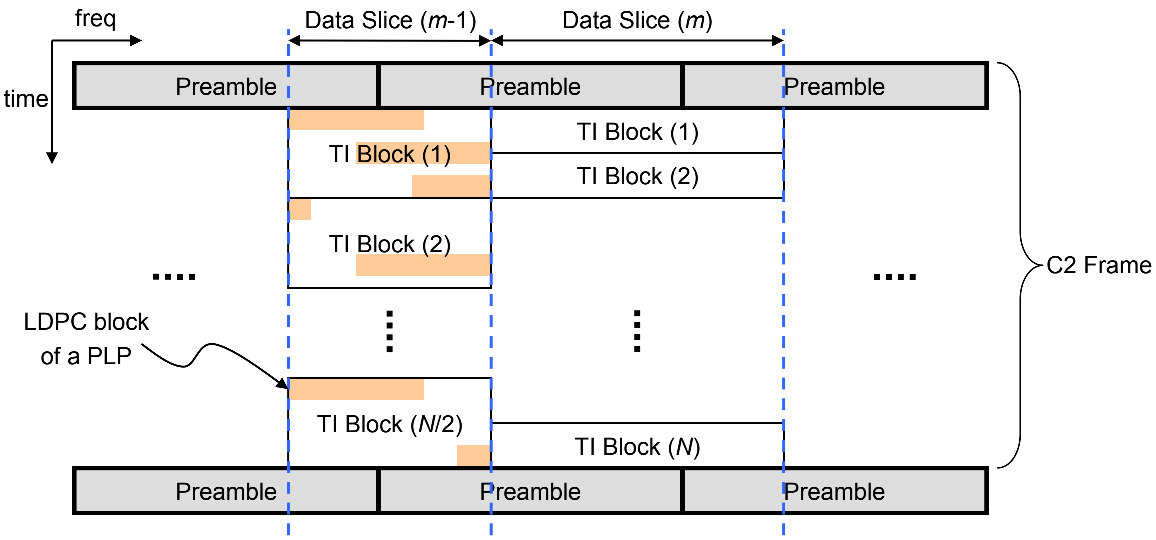


Figure 53: Typical example of time interleaving for the Data Slice

The rectangular blocks within the Data Slice are TI blocks, the basic unit of time interleaving in the Data Slice. One can see the boundary of the FEC blocks are aligned with neither C2 frame nor TI block. This means no integer number of FEC blocks are guaranteed within each TI block or C2 frame. In addition, the TI block may contain the FEC blocks of different PLPs. However, one C2 frame should contain integer number of TI blocks.

The time interleaving is applied for Data Slice basis rather than individual PLP inside. The depth of time interleaving (m-1) Data Slice is different from that of m Data Slice, as shown in Figure 53. This seems reasonable since the burst noise characteristic is considered as same per Data Slice, not per PLP. In addition, it reduces the buffering complexity otherwise required for each PLP.

For the case of L1 signalling part2, Figure 54 shows a typical example of the time interleaving structure. It shows two preambles of the two C2 frames, each of which has different preamble structure. Each preamble is composed of more than one L1 TI blocks. The preceding preamble contains the narrower L1 TI blocks and the following the wider. As a consequence, the number of repeated L1 TI blocks is different from each other.

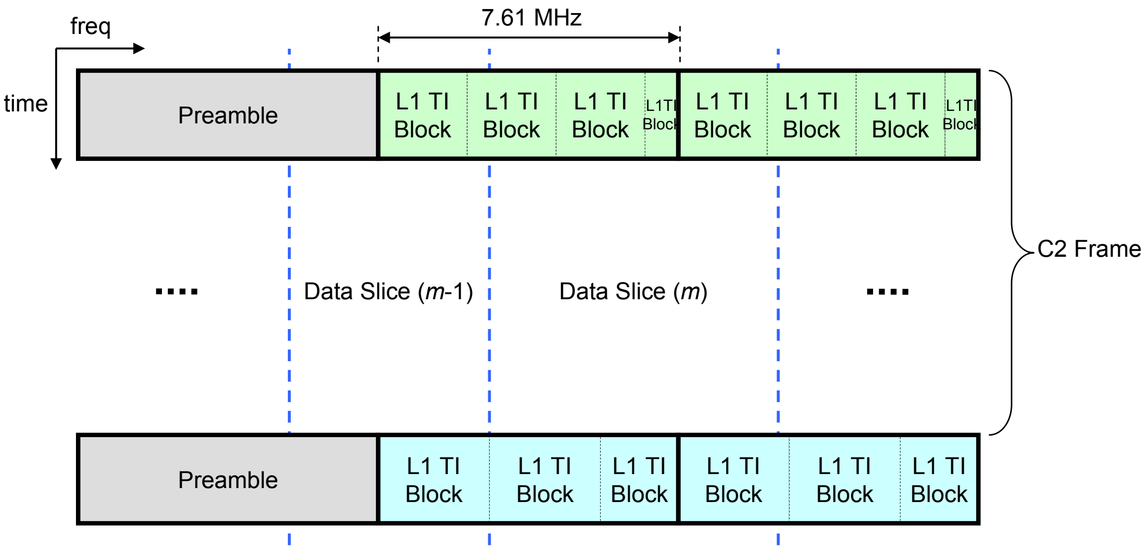


Figure 54: Typical example of time interleaving for the L1 signalling part2

L1 TI block is the basic unit of time interleaving for the L1 signalling part2. The L1 TI block is unique in each preamble: the whole L1 part2 information should be included in a single L1 TI block. However, repetition of the L1 TI block is permitted to fulfil the entire channel raster bandwidth (7.61 MHz in this example). As a consequence, the L1 TI block should contain the integer number of FEC blocks.

The detailed time interleaving structure for the Data Slice is illustrated in Figure 55. This is a kind of block time interleaver where the input cells are written in diagonal direction and outputs are read in row direction. The different coloured cells in the figure indicate the cells allocated to different OFDM symbols unless time interleaving is applied. These cells are originally output as adjacent cells but spread both in time and frequency after time interleaving.

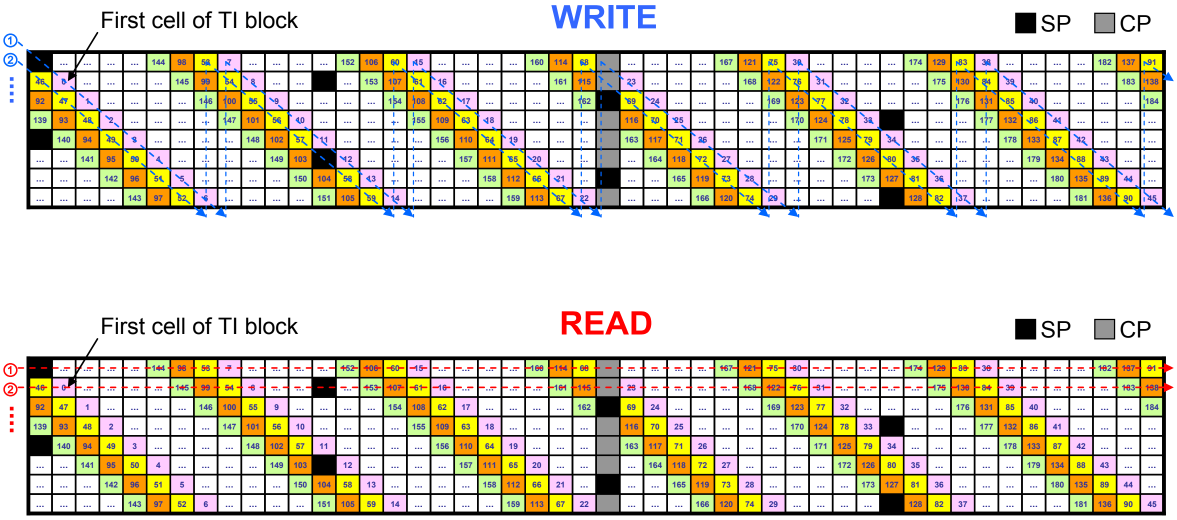


Figure 55: Time interleaving structure for the Data Slice

The number of columns is equal to the width of the Data Slice. The number of rows is same as time interleaving depth which can be one of four options: 1, 4, 8 or 16 OFDM symbols. The depth of '1' means no time interleaving. One should note that the pilot positions are considered but excluded in the interleaving sequence. For example, when the sequence meets pilot position for the input cell, the pilot positions should be skipped and jumped to the next sequence to write the input cell. This rule applies for both writing and reading process. The maximum value of the TI block width and TI depth is 3,408 carriers and 16 OFDM symbols respectively. Thus, the interleaving memory of maximum 54,424 cells is required.

Almost same interleaving sequence is applied also for the interleaving L1 signalling part2 data. Figure 56 shows the typical example of time interleaving structure of L1 part2.

The relevant process before and after time interleaving includes:

(1) dividing L1 signalling part2 data into sub-blocks and encoding them into separate FEC blocks
(2) time interleaving if the option is chosen
(3) inserting L1 header to indicate the size of L1 signalling part2 data and the time interleaving parameters
(4) repeating header and L1 TI block to fulfil the entire channel raster bandwidth

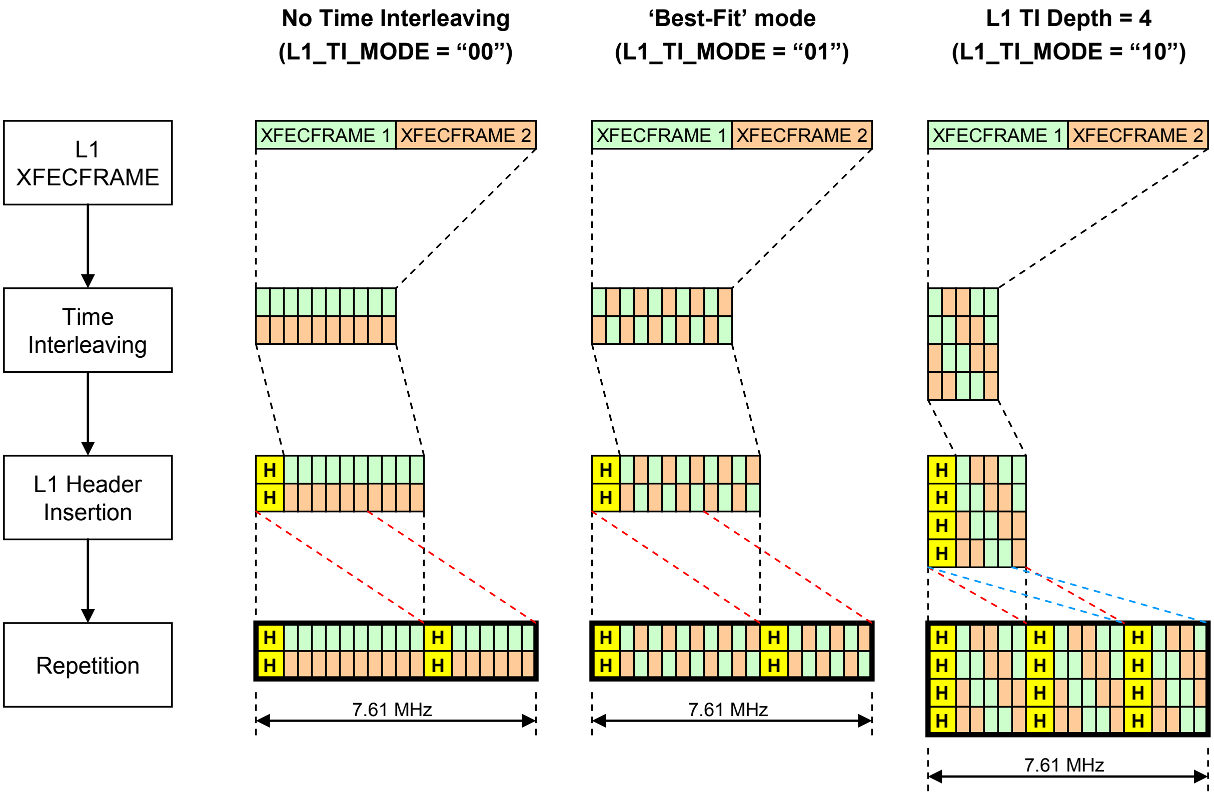


Figure 56: Time interleaving structure for the L1 signalling part2

As for the time interleaving process, writing diagonal-wise and reading row-wise the L1 part2 data cells are same as those of Data Slice time interleaving. The only difference is that no pilot positions are considered in the interleaving sequence as shown in the figure. The number of rows is same as time interleaving depth and there are four options: no TI, best-fit, 4 or 8 OFDM symbols. The option 'best-fit' means the depth is equal to the number of FEC blocks used for encoding L1 signalling part2 information. The number of columns is determined by the number of encoded L1 part2 data cells divided by the interleaving depth described above. The maximum number of L1 signalling part2 data bits is 32,766 bits so the interleaving memory of maximum 16,383 cells is required.

As for the 'best-fit' option, no additional overhead is required compared to no time interleaving. The remained two options of depth '4' or '8' are used when the amount of L1 part2 information is so small that, for example, even single OFDM symbol is enough to carry them. In this case, the lack of interleaving robustness may be a problem according to the channel status. It's serious since the L1 signalling part2 information has highest priority of protection and is crucial for decoding the service data stream in the receiver. On the contrary, the Data Slice is free from this problem since it can freely select up to 16 OFDM symbols of time interleaving depth regardless of the bit rate. To avoid this situation, the transmitter can arbitrarily select the time interleaving depth larger than that of 'best-fit' case. This is illustrated in Figure 56 where the third option (TI depth of 4 OFDM symbols) provides better interleaving robustness compared to second option (best-fit) for the same amount of L1 signalling part2 data (20 cells in the figure).

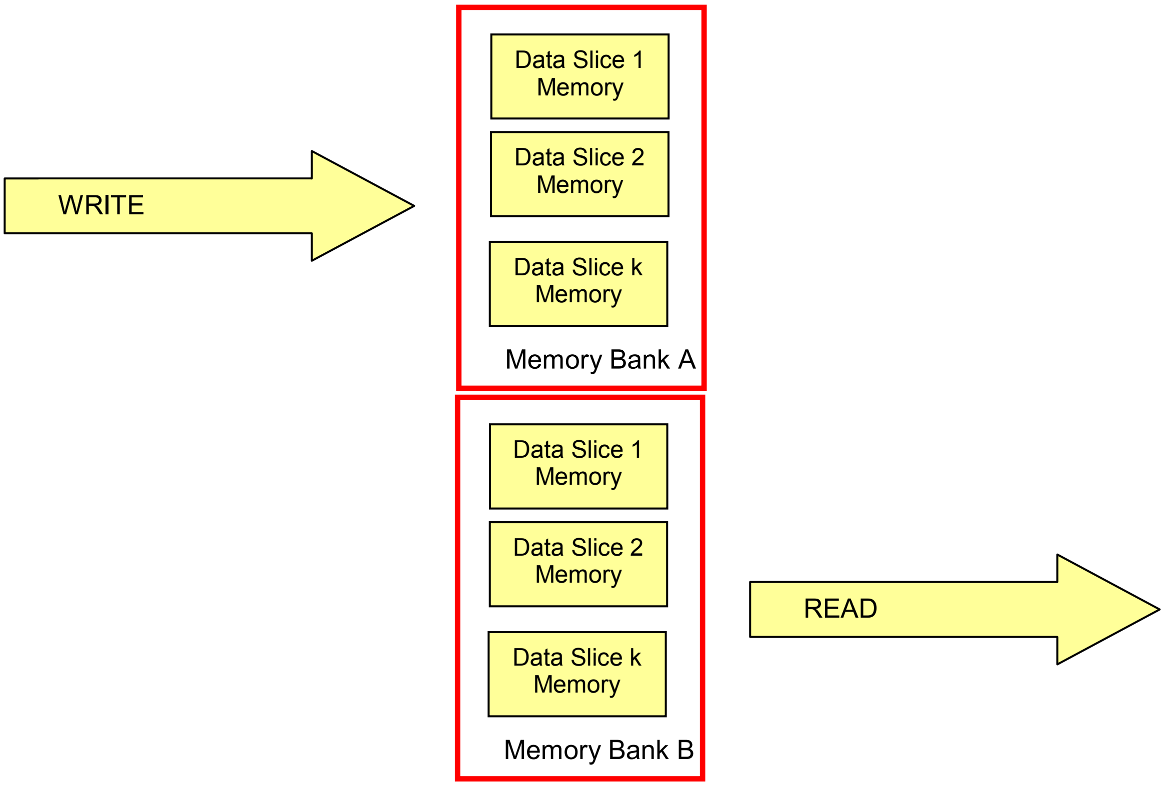


Figure 57: Time interleaver memory in the modulator side

Simple example of the TI memory structure in the modulator side is depicted in Figure 57. This is so-called 'ping-pong' architecture which comprises two separate memory banks. Each memory bank may be composed of several memory blocks when multiple Data Slices exist within the C2 system. Whilst writing the input cells to one memory bank, the transmitter read out the interleaved sequence from the other memory bank and vice versa on next turn.

The same architecture could be used for the receiver side. However, it requires the double size of the interleaving memory because of ping-pong scheme. This architecture definitely raises the cost of the receiver. To reduce the implementation cost, more efficient implementation of time de-interleaver is shown in Figure 58.

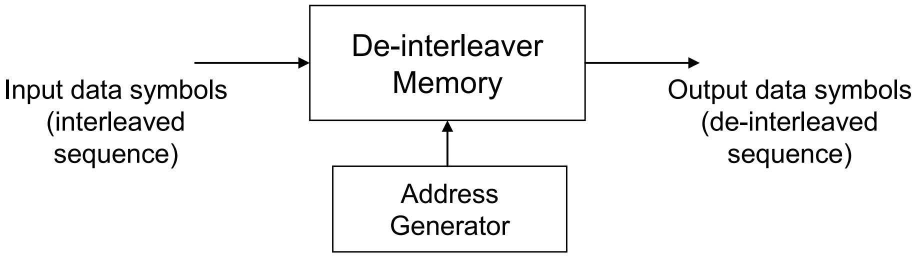


Figure 58: Efficient implementation of the time de-interleaver

The de-interleaver consists of single buffer memory and simple address generator. The size of de-interleaver memory shown in Figure 58 is exactly identical to one memory bank size shown in Figure 57. So it reduces the implementation complexity by around a half.

An intuitive example of address generation is illustrated in Figure 59. The interleaving rule is basically 'in-placement' algorithm: the same addresses used for reading output cells of one block are used for writing the input cells of next block. By doing this, need of double buffering as in Figure 58 is absolutely eliminated.

The de-interleaving sequence is the inverse of that of the time interleaving shown in Figure 55. That is, the input cells are written in row direction and output cells are read in diagonal direction. In Figure 59, the first TI block is written row-wise which is indicated as blue path. After writing all input cells of the first TI block, the de-interleaving starts by reading the cells of the first TI block in diagonal direction (red path). At the same time, the input cell of the second TI block is written into the same address immediately after reading the output cell. This process continues until complete first TI block is read out following de-interleaving sequence and complete second TI block is written, both along the red path. The same concept of memory access is applied for reading the de-interleaved second TI block and writing the third input TI block, of which path is indicated as green colour.

This addressing seems complex but can be easily generated with the following rule. The reading address can be generated by circularly shifting the writing address in column direction. The amount of this circular shifting is dependent on the column and it increases as the column index increase. When the address or shift value reaches its maximum, it returns back to its initial value, which is 'circular' operation.

For example, the first column has no such a shift: the shift value starts from '0'. For the second column, the shift value is increased to '1' so the red path (first sequence) address is shifted down by the shift value '1' from the blue path (second sequence). The same rule is true for between blue and green path. For the third column, the shift value is increased to '2', and so on. The 'circular' operation can be found on the fifth column where the shift value is '4' and the circular shift folds the address of green path to the start of the column, or red path one. The above description is expressed as following equations:

$$
\begin{aligned}
c_{i,j} &= i \; mod \; C \\
s_{i,j} &= (j \times c_{i,j}) \; mod \; R \\
r_{i,j} &= [s_{i,j} + (i \; div \; C)] \; mod \; R
\end{aligned}
\qquad \text{Eq. 7.1}
$$

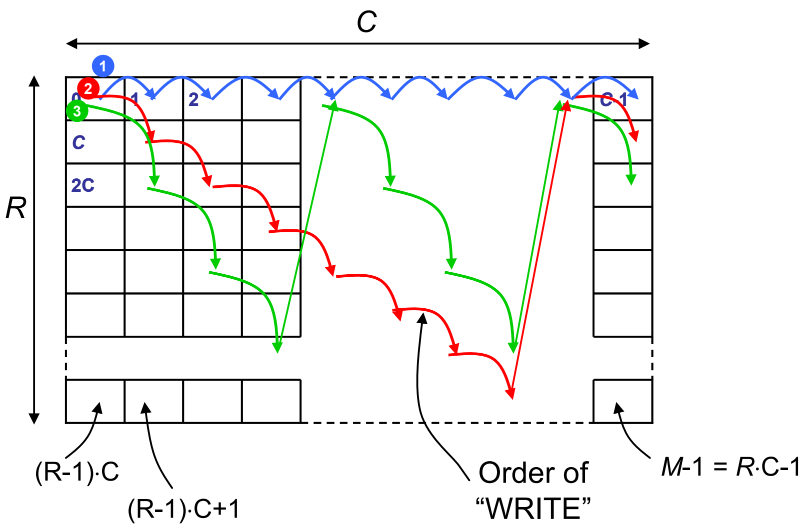


Figure 59: Example of address generation for the single buffer implementation

where ci,j, ri,j and Si,j are the column address, row address and shift value respectively for writing the i-th cell of the j-th TI block. According to the description above, the same address is used for reading out the i-th cell of the (j-1)-th TI block. The constant R and C are the number of rows and columns respectively. The operator mod and div are defined as:

$$x \; div \; y = \left\lfloor \frac{x}{y} \right\rfloor, \qquad x \; mod \; y = x - y \left\lfloor \frac{x}{y} \right\rfloor \qquad \text{Eq. 7.2}$$

If a linear array memory is used for the de-interleaving memory in Figure 58, one-dimensional address Li,j for the i-th cell of the j-th TI block is obtained by simple conversion from 2-dimensional coordinates (ci,j, ri,j):

$$L_{i,j} = C \cdot r_{i,j} + c_{i,j} \qquad \text{Eq. 7.3}$$

In conclusion, DVB-C2 system provides powerful time interleaving functionality to combat any type of burst error at the cost of reasonable complexity. The interleaving robustness is controlled by Data Slice and/or C2 frame basis (for L1 signalling part2). The time interleaving depth varies from 1 to 16 OFDM symbols for the Data Slice and arbitrary depth of up-to 8 OFDM symbols is possible for the L1 signalling part2. This provides sufficient flexibility to adapt to any given cable channel condition, which is expected to help successful worldwide deployment of C2 system.

8 Frequency Interleaving

The frequency interleaver in DVB-C2 is applied to the payload cells of a given Data Slice from one OFDM symbol to the next. The effect of the frequency interleaver is to shuffle the sub-carrier locations of the payload cells of a given Data Slice in a pseudo-random manner from OFDM symbol to symbol. The frequency interleaver combats two possible channel effects especially in the case when a Data Slice carries more than one PLP:

(1) Channel selectivity – multipath propagation (admittedly low in a cable environment) causes frequency selectivity on the channel. This means that some parts of the spectrum fade more than others. Multipath propagation in cable transmissions is caused by reflections at cable joints and splits which are static. This produces a static frequency selectivity pattern in which the parts of the spectrum are on average faded and others are boosted. Frequency interleaving avoids the situation in which the payload cells of a given PLP are consistently mapped to the faded part of the spectrum. By pseudo-randomly changing the cell to OFDM sub-carrier position mapping from symbol to symbol, the probability of a given cell being mapped to a faded sub-carrier becomes similar for all cells within the data-slice.

(2) Narrow-band interference – the DVB-C2 specification has provision for designating frequency notches in the spectrum for known narrow-band interferers. The presence of unknown or intermittent narrow-band interferers would cause degradation to the payload cells allocated to sub-carriers whose positions coincide with that of the interferers. If such interferers are sufficiently persistent, they may disproportionately affect the payload cells of a particular PLP. The pseudo-random allocation of sub-carrier locations to the payload cells of PLPs from OFDM symbol to symbol avoids this scenario.

8.1 Design of the Frequency Interleaver

The frequency interleaver is made up of two main blocks as illustrated in Figure 60 i.e. the grey shaded blocks. In its basic operation, the N_{DS} payload cells of each OFDM symbol are written sequentially into an interleaver RAM. Once all N_{DS} payload cells are written, interleaving is performed by reading out the contents of the interleaving RAM using a series of read addresses generated in a pseudo-random manner by the address generator block. The rationale for the FIFO and RAM Control blocks will be explained later.

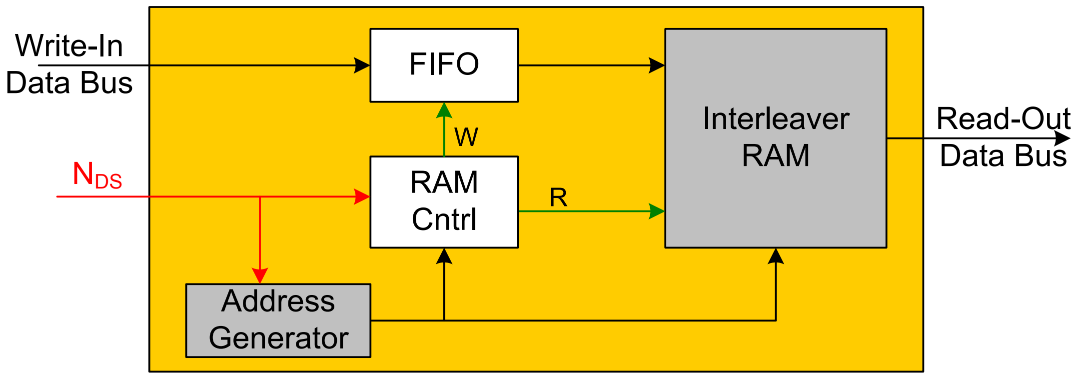


Figure 60: Architecture of frequency (de)-interleaver

8.2 Pseudo-random address generation

The address generator must be capable of generating enough addresses to read out all the contents of the RAM. This means that, the maximum number of addresses that the generator is capable of should be equal to the maximum possible number of payload cells in one Data Slice. According to the specification, a Data Slice can be comprised of a maximum of:

$$N_{max} = (K_{DS,max} - K_{DS,min}) - N_{SP,D}x_{=24}, \qquad \text{Eq. 8.1}$$

where $(K_{DS,max} - K_{DS,min}) \leq 3408$ and $N_{SP,Dx=24}$ is the number of scattered pilots in 3408 sub-carriers for the $D_x = 24$ scattered pilot pattern and N_{CP} is the number of continual pilots in 3408 sub-carriers.

The pseudo-random addresses are generated by means of a maximal-length pseudo-random binary sequence (PRBS) generator illustrated in Figure 61.

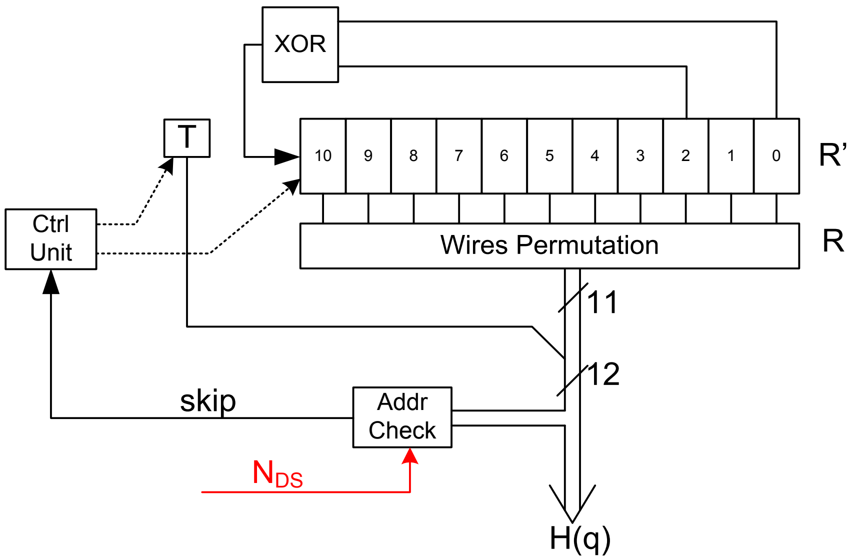


Figure 61: Pseudo-random address generator

The address generator itself is made up of three main parts:

(1) A maximal-length PRBS generator implemented in the form of a feedback shift register. The generator polynomial of this PRBS is:

$$R'_i[10] = R'_{i-1}[0] \oplus R'_{i-1}[2] \qquad \text{Eq. 8.2}$$

This means that the MSB (bit 10) is updated after each cyclic shift using XOR of the LSB (bit 0) and bit 2. As the shift register has 11 bits, it can generate a maximal-length sequence of pow(2, 11) = 2048 unique values before repeating.

(2) There is a wires permutation circuit which takes the 11 bits of each value generated by the PRBS and shuffles them in a pre-determined manner to create the 11 least significant bits of the pseudo-random interleaver RAM addresses. DVB-C2 frequency interleaving employs two wires permutation circuits – one for even and another for odd OFDM symbols. Each wires permutation was designed using a directed search process that ensures the most optimum interleaving performance. The wires

permutations can be described by Table 10. Figure 62 illustrates the how the circuit maps the input bits R' to the output bits R for H_0.

R'_i bit positions	10	9	8	7	6	5	4	3	2	1	0
R_i bit positions (H_0)	7	10	5	8	1	2	4	9	0	3	6
R_i bit positions (H_1)	6	2	7	10	8	0	3	4	1	9	5

Table 10: Wire permutations

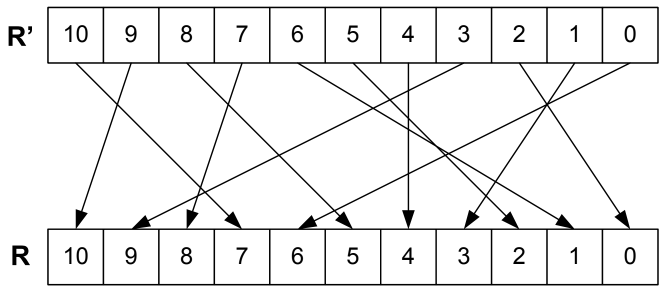


Figure 62: Wires permutation execution for Ho

(3) Valid address control circuitry comprising a toggle circuit which appends the MSB to the address generation to form a complete 12 bit address. For even addresses generated by the PRBS and wires permutation circuit, the toggle sets this MSB to 0 whilst for odd addresses it sets the MSB to 1. The control and address check blocks ensure that any generated address is valid i.e. it lies within the range of 0 to $N_{DS} - 1$. Otherwise, the address is discarded and the shift register clocked again to generate a new address.

In general the pseudo-random address *H(q)* generation process can be described by the following algorithm:

q = 0;
for (i = 0; i < 4096; i = i + 1)
{ $H(q) = (i \bmod 2).2^{11} + \sum_{j=0}^{10} R_i(j).2^j;$ Eq. 8.3
if (H(q)<NDS(n)) q = q+1; }

8.3 Frequency Interleaver implementation

The number of payload cells per Data Slice per OFDM symbol (N_{DS}) can vary from symbol to symbol. For preamble symbols, the maximum number of cells to go through the frequency interleaver is 2840 cells – for preamble symbols therefore, it can be said that N_{DS} = 2840. For data symbols, N_{DS} comprises the number of cells between ($K_{DS,max}$ - $K_{DS,min}$) minus the number of continual pilots, scattered pilots, reserved tones and cells that are located in notches within the Data Slice. Most of this information is carried in the Layer 1 signalling. For this purpose, assume that a function DataCells (*slice number, symbol number, L1 info*) exists in the modulator to provide N_{DS} for a given Data Slice number, symbol number and from the

Layer 1 information. The frequency interleaver must be capable of interleaving payload cells of the largest possible Data Slice with max(N_{DS}) cells given that ($K_{DS,max}$ - $K_{DS,min}$) ≤ 3408. This means that the interleaver must be capable of dealing with N_{max} payload cells where:

$$N_{max} = (K_{DS,max} - K_{DS,min}) - N_{SP,Dx=24} - N_{CP} \qquad \text{Eq. 8.4}$$

and $N_{SP,Dx=24}$ is the number of scattered pilots in 3408 sub-carriers for the D_x = 24 scattered pilot pattern and N_{CP} is the number of continual pilots in 3408 sub-carriers. Note that N_{max} > 2840 so the interleaver memory is enough even for the preamble symbols. Designate the payload cells of symbol *n* to be interleaved as $X_{n,q}$ where *q* = 0, 1, . . N_{DS} for a given slice and let the equivalent interleaved cells be designated as $A_{n,q}$.

The frequency interleaver memory is made up of two banks: Bank A for even OFDM symbols and Bank B for odd OFDM symbols. Each memory bank comprises of N_{max} locations. DVB-C2 uses odd-only pseudo-random interleaving. In this the payload cells from even symbols (symbol number of form 2*n*) of the Data Slice are written into the interleaver memory Bank A in a sequential order and read out in a permuted order determined by $H_0(q)$. Similarly, payload cells from odd OFDM symbols (symbol number of form 2*n*+1) of the Data Slice are written into interleaver memory Bank B in a sequential order and read out in a permuted order determined by $H_1(q)$. In each case, the permuted order addresses $H_{[0,1]}(q)$ are provided by the pseudo-random address generator of Figure 61 with the appropriate wires permutation. In order to latency through the frequency interleaver, when Bank A is being written (incoming even symbol), Bank B is also being read (outgoing previous odd symbol) at the same time. Indeed, the sequential counter *q* doubles as both the sequential write address and the look up table index to each of the permutation functions $H_{[0,1]}(q)$.

If all symbols in the Data Slice contained $N_{max} = C_{data}$ payload cells, then the number of write addresses for symbol number 2*n*+1 would match the number of read addresses for symbol number 2*n* otherwise some data cells of symbol 2*n* will be skipped. Unfortunately N_{DS} can be different from symbol to symbol. Suppose $N_{DS}(2n) < N_{DS}(2n+1)$ then the pseudo-random address generator $H_0(q)$ would have to produce more addresses than there are cells to be read from memory Bank A because the sequential write address counter q for Bank B would range from 0 to $N_{DS}(2n+1)-1$ ($>N_{DS}(2n)$). The case in which $N_{DS}(2n) > N_{DS}(2n+1)$ can also occur. In this case the sequential write address counter for Bank B would need to exceed $N_{DS}(2n+1)-1$ as more $H_0(q)$ read addresses are needed for Bank A. Recalling that the function DataCells (*slice number, symbol number, L1 info*) returns the number of payload cells in the current slice for the given symbol and noting that HoldBuffer is a small amount of storage with write address *wptr* and read address *rptr* (this represents the FIFO in Figure 60), the interleaving proceeds as follows at the start of any even symbol of number 2*n*:

1. $q = 0$;
2. C_{max} = **max**(**DataCells**(*slice number* , 2*n-1*, *L1 info*), **DataCells**(*slice number* , 2*n*, *L1 info*))
3. Generate address $H_1(q)$;
4. rdEnable = ($H_1(q)$ < **DataCells**(*slice number* , 2*n-1*, *L1 info*));
5. wrEnable = (*q* < **DataCells**(*slice number* , 2*n*, *L1 info*));
6. if (rdEnable) Read cell $A_{2n-1,q}$ from location $H_1(q)$ of memory Bank B;

7. Store cell $X_{2n,q}$ into location *wptr* of HoldBuffer and increment *wptr;*
8. if (wrEnable):
    a. Write cell *rptr* of HoldBuffer into location *q* of memory Bank A and increment *rptr*.
    b. If(*wptr* == *rptr*) reset both *rptr* = *wptr* = 0.
9. Increment *q*;
10. if ($q < C_{\max}$) goto 3

Then with symbol 2*n*+1 at the input of the interleaver:

1. $q = 0$;
2. C_{max} = **max**(**DataCells**(*slice number* , 2*n*, *L1 info*), **DataCells**(*slice number* , 2*n+1*, *L1 info*))
3. Generate address $H_0(q)$;
4. rdEnable = ($H_0(q)$ < **DataCells**(*slice number* , 2*n*, *L1 info*));
5. wrEnable = (*q* < **DataCells**(*slice number* , 2*n+1*, *L1 info*));
6. if (rdEnable) Read cell $A_{2n,q}$ from from location $H_0(q)$ of memory Bank A;
7. Store cell *q* of $X_{2n+1,q}$ into location *wptr* of HoldBuffer and increment *wptr;*
8. if (wrEnable):
    a. Write cell *rptr* of HoldBuffer into location *q* of memory Bank B and increment *rptr*.
    b. If(*wptr* == *rptr*) reset both *rptr* = *wptr* = 0.
9. Increment *q*;
10. if ($q < C_{\max}$) goto 3

The required width for each memory location depends on the resolution with which each cell is represented after QAM mapping.

Care should be taken to implement the interleaving function in the correct sense. As shown in the steps detailed above, for each symbol, the interleaver should write to the memory in sequential order and read in permuted order.

8.4 Frequency De-interleaving

The frequency de-interleaver in DVB-C2 is applied to the received payload cells of a given Data Slice from one OFDM symbol to the next. The de-interleaver has the same requirement for 2 Banks of memory each comprising of $N_{\max}$ locations. The width of each memory location may be different from that at the interleaver.

Designate the received and equalised payload cells of symbol *n* to be de-interleaved as $A'_{n,q}$ where $q = 0, 1, . . N_{DS}$ for a given slice and let the equivalent de-interleaved cells be designated as $X'_{n,q}$. Recalling that the function DataCells (*slice number, symbol number, L1 info*) returns the number of payload cells in the current slice for the given symbol and noting that HoldBuffer is a small amount of storage with write address *wptr* and read address *rptr*, the de-interleaving proceeds as follows at the start of even symbol number 2*n*:

1. $q = 0$;
2. C_{max} = **max**(**DataCells**(*slice number* , 2*n-1*, *L1 info*), **DataCells**(*slice number* , 2*n*, *L1 info*))
3. Generate address $H_0(q)$;
4. rdEnable = (q < **DataCells**(*slice number* , 2*n-1*, *L1 info*));
5. wrEnable = ($H_0(q)$ < **DataCells**(*slice number* , 2*n*, *L1 info*));
6. if (rdEnable) Read cell $X'_{2n-1,q}$ from location q of memory Bank B;
7. Store cell $A'_{2n,q}$ into location *wptr* of HoldBuffer and increment *wptr;*
8. if (wrEnable):
    a. Write cell *rptr* of HoldBuffer to location $H_0(q)$ of Bank A and increment *rptr*.
    b. If(*wptr* == *rptr*) reset both *rptr* = *wptr* = 0.
9. Increment q;
10. if ($q < C_{\max}$) goto 3

Then with symbol $A'_{2n+1,q}$ at the input of the de-interleaver:

1. $q = 0$;
2. C_{max} = **max**(**DataCells**(*slice number* , 2*n*, *L1 info*), **DataCells**(*slice number* , 2*n+1*, *L1 info*))
3. Generate address $H_1(q)$;
4. rdEnable = (q < **DataCells**(*slice number* , 2*n*, *L1 info*));
5. wrEnable = ($H_1(q)$< **DataCells**(*slice number* , 2*n+1*, *L1 info*));
6. if (rdEnable) Read cell $X'_{2n,q}$ from location q of memory Bank A;
7. Store cell $A'_{2n+1,q}$ into location *wptr* of HoldBuffer and increment *wptr;*
8. if (wrEnable):
    a. Write cell *rptr* of HoldBuffer to location $H_1(q)$ of Bank B and increment *rptr*.
    b. If(*wptr* == *rptr*) reset both *rptr* = *wptr* = 0.
9. Increment q;
10. if ($q < C_{\max}$) goto 3

The required width for each memory location depends on the resolution with which each cell is represented after channel equalisation. Each de-interleaver memory cell would need to hold at least the complex cell information and the channel state information for the cell.

9 Usage of Pilots

Usage of pilots is a receiver design issue, and there is no enforcement on how to use them. However, it is also true that pilots are inserted with some degree of assumption on how to they will be used. This section provides an introductory example of pilot usage, and will hopefully help explain the reasons why DVB-C2 has such a pilot structure.

Pilot Signalling
In general, pilot signalling involves some kind of signalling that is known to the receiver. The reason for signalling known data is to make channel estimation and/or synchronisation easier and more reliable.

Pilots may be inserted in specific time periods (time division manner) or they may be inserted in specific frequency bands (frequency division manner). In another case, they may be inserted using a certain code (code division manner), or even a combination of a number of these methods. In the DVB-C2 system, pilots are inserted in frequency and time division manner. In other words, they are inserted in particular subcarriers within the OFDM symbol, and the pilot subcarriers are changed from one symbol to the next.

Since the pilots are known to the receiver, they don't carry any information at all. Hence it might look like a waste of spectrum, but carefully designed pilots make receivers much simpler and much more reliable with very low spectral penalty. In that sense, having pilots are a good way to consume spectrum.

Purpose of Pilots
In the DVB-C2 system, there are predominantly two reasons for the insertion of pilots:

- Channel estimation (and subsequent equalisation)
- Synchronisation

The details of pilot usage are discussed in the subsequent sections.

9.1 Pilot structure

DVB-C2 has two types of symbols within a frame. At the begging of the frame there are preamble symbol(s), carrying signalling information. The other symbol type is data symbols, which come after the preamble symbols and carry actual data. Each type of symbol has a different pilot structure. The preamble symbols have preamble pilots which are inserted every 6 subcarriers in the frequency axis, and the pilot carrier location doesn't change with time. The preamble pilots are inserted primarily for frame synchronisation, channel estimation and synchronisation in order to decode the preamble data reliably.

Data symbols have three kinds of pilots. The first type is scattered pilots (SP) which are inserted every 48 or 96 subcarriers depending upon the chosen Guard Interval (GI) fraction. The pilot subcarrier location is shifted by 12 or 24 subcarriers from symbol to symbol. As result, scattered pilots appear every 4 symbols in a particular subcarrier. The amount of pilot location shift is defined as 'Dx', which is 12 or 24 subcarriers, while the number of symbols separation between pilots is defined as 'Dy', which is always 4. Hence the carrier separation between consecutive pilots can be defined as 'Dx*Dy'. The second pilot type is continual

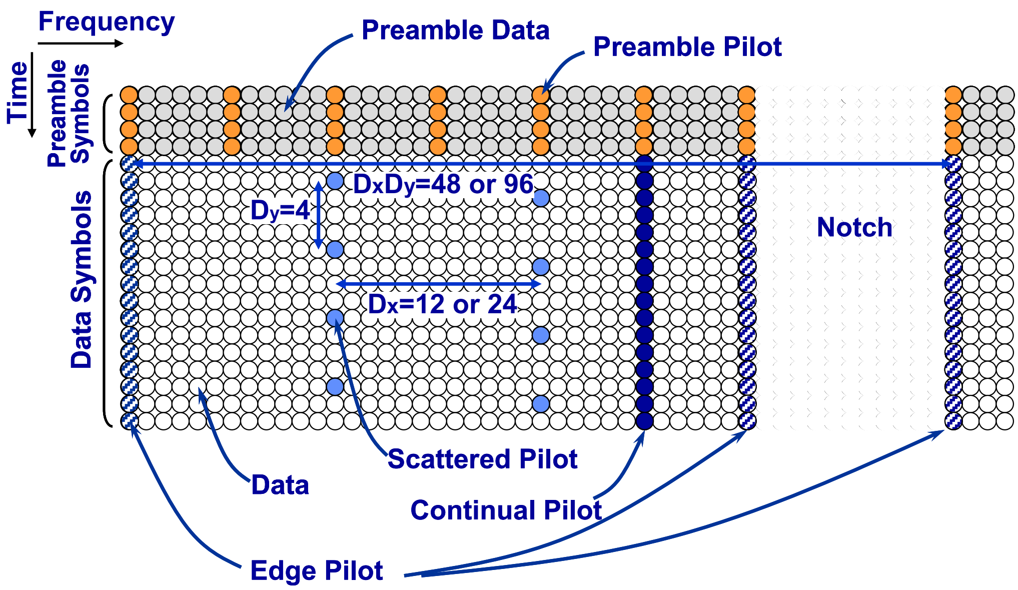


Figure 63: Pilot structure in time-frequency diagram

pilots (CP), which are inserted continually in time using predefined pseudo-random carrier locations. The third pilot type is edge pilots (EP), which are inserted at the lowest and highest frequency subcarriers in bundled OFDM carriers. If there is a notch, EPs are also inserted at both edges of the notch. SPs are inserted for channel estimation, while CPs are inserted for synchronisation. EPs are primarily for channel estimation, but they can be used for synchronisation as well.

9.2 Channel estimation & equalisation for OFDM system

Since there is a cyclic prefix, GI, it is possible to extract a block of data which appears as though the block of data is repeated in time (assuming the channel impulse response is shorter than the GI period). Taking the FFT of such a block of data, we get the frequency domain received data, which can be given by the following equation.

$$y_{k,l} = H_{k,l}\, x_{k,l} + n_{k,l} \qquad \text{Eq. 9.1}$$

Where $y_{k,l}$ is the received data at symbol index l, carrier index k, $H_{k,l}$ is the frequency domain channel impulse response, $x_{k,l}$ is the transmitted data and n is the noise. The channel estimation process estimates $H_{k,l}$, so we can derive $H_{k,l}$ by

$$H_{k,l} = y_{k,l} / x_{k,l} \qquad \text{Eq. 9.2}$$

We can then use the estimated channel response $H_{k,l}$ to equalise the received data y by

$$x_{k,l} = y_{k,l} / H_{k,l} \qquad \text{Eq. 9.3}$$

Unfortunately, the transmitted data $x_{k,l}$ is known only at the pilot subcarriers, so the above approach is only applicable for these pilot subcarriers. What we can do for the rest of the subcarriers is employ interpolation between these estimated channel responses on the pilot subcarriers. There are many interpolation methods proposed, and depending upon the method, the character of the channel estimation varies. In the following sections we are going to discuss two typical interpolation methods, namely 'time and frequency interpolation' and 'frequency only interpolation'.

9.2.1 Time and frequency interpolation

For this method, firstly we interpolate in the time axis on scattered pilot bearing subcarriers (which are every Dx subcarriers). The interpolation order is Dy. Secondly, we interpolate in the frequency axis to get the channel responses for all subcarriers. The frequency axis interpolation order is Dx since the time axis interpolations are carried out first.

These interpolation orders are an important factor in order to decide the character of the interpolation method. The time axis interpolation order decides Doppler frequency capability, and the Nyquist limit for Doppler frequency is given by $1/T_{sym}/Dy/2$, where T_{sym} is OFDM symbol period. On the other hand, the frequency axis interpolation order decides echo delay capability, and the Nyquist limit for the echo delay is given by Tu/Dx, where Tu is the OFDM useful symbol period. The actual Nyquist limit for DVB-C2 is 37.3 or 18.7usec depending upon the SP pattern. GI periods for these scattered pilot patterns are 7 and 3.5usec respectively, hence the GI period to Nyquist limit ratio is around 0.18, which is considerably small. This small ratio can be exploited for reducing noise in the channel estimation and/or relaxing the interpolation filter requirements. Later one might be more important for DVB-C2 system, since receiver need to support very high QAM size and code rate such as 4096 QAM with code rate 9/10 hence the interpolation requirement originally should be very high.

Another important issue regarding this approach is that the time interpolation could introduce delay into the channel estimation, which requires extra memory to delay the main data symbols in order to align them with the channel estimation for equalisation. Generally speaking, a more accurate time interpolator requires a higher filter order, which introduces more delay and results in larger delay memory requirement. Considering cable environments, it is unlikely that we will require a high order temporal interpolation filter, hence it is possible to use just linear interpolation, which only requires a three symbol delay. Potentially, it might be possible to use some form of extrapolation technique which doesn't require any delay.

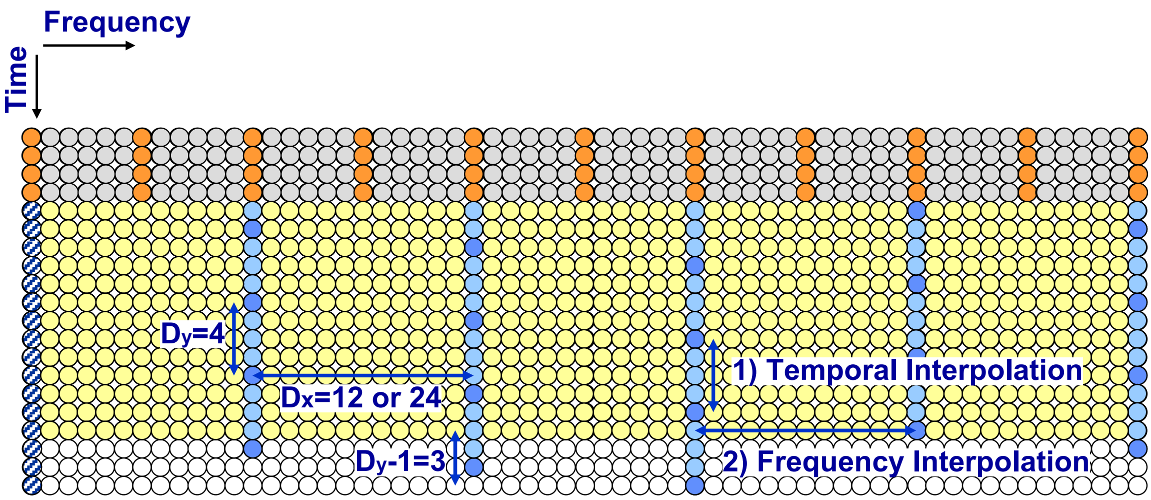


Figure 64: Pilot structure used for interpolation in time and frequency domain

9.2.2 Frequency-only interpolation

For the frequency only interpolation case, the interpolation is simply carried out in the frequency axis only. Because there is no time interpolation, there is no delay for the channel estimation, hence it does not require any delay memory. Another merit of this technique is that the equaliser automatically cancels its common phase error (CPE), since the channel estimation is independent from past and future symbols.

The downside of this approach is that it has a very high upsample rate for the frequency interpolation. The interpolation order is 48 or 96 (depending upon GI fraction), resulting in a limited echo delay capability. The Nyquist limit of the echo delay equates to 9.3 or 4.7usec, which gives a GI period to Nyquist limit ratio of around 0.75. This is still a manageable ratio, but there is not much room to apply noise reduction to the channel estimation or interpolation filter order reduction.

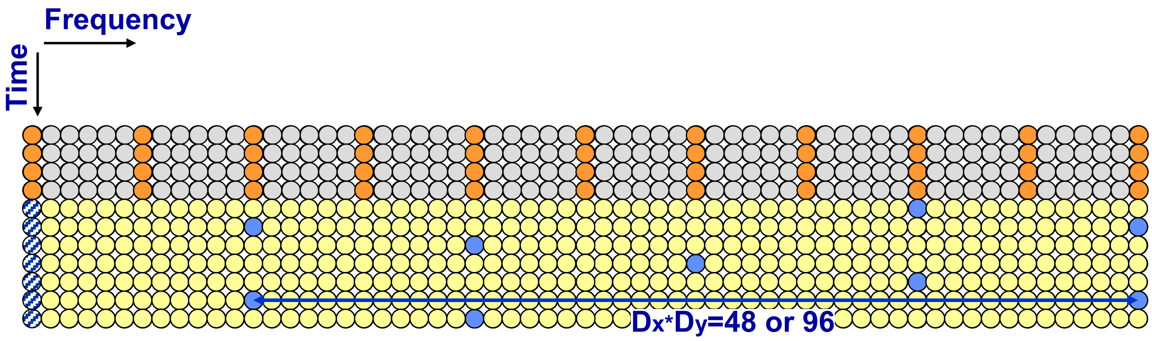


Figure 65: Pilot structure used for interpolation in frequency domain only

9.3 Synchronisation

Receivers always need to synchronise a number of parameters to the transmitter. In the DVB-C2 case, it is important to synchronise carrier frequency, sampling frequency and symbol timing. Carrier frequency offset and sample frequency offset cause inter-carrier-interference (ICI), resulting in a significant level of decoding errors. Appropriate symbol timing alignment is required, otherwise the demodulation results suffer from not just inter-symbol-interfere (ISI) but also inter-carrier-interfere (ICI), since the symbol timing error causes incomplete OFDM symbols that generate ICI.

This section will also discuss common-phase-error (CPE) estimation. Strictly speaking this may not be a synchronisation issue, but it uses a similar technique to the one used for the carrier frequency synchronisation, hence we will discuss it here.

9.3.1 Coarse AFC

AFC stands for Automatic Frequency Control, i.e. carrier frequency synchronisation. In most OFDM systems, the carrier frequency synchronisation is split into two stages, coarse and fine. Coarse AFC is Automatic Frequency Control with subcarrier resolution. The other one is fine AFC, which is carrier frequency synchronisation within subcarrier spacing. This will be discussed in a later section.

The basic approach for coarse AFC for DVB-C2 is detecting pilot carrier(s) that have a unique pattern (modulation phase or location) relative to its carrier frequency (RF). Since receivers know which carrier frequency to tune to, it is possible to pre-calculate the unique pilot pattern and coarse AFC can be carried out by searching for the pattern in the received data. The DVB-C2 system has two kinds of pilots useful for coarse AFC. One is the preamble pilots and the other is the continual pilots. Both cases are going to be explained in the following sections.

Using preamble pilots

Preamble pilots are differentially modulated between consecutive pilots in the frequency direction with a PRBS sequence that is unique to the carrier frequency. We can use this sequence for the coarse AFC. The procedure would be as follows:

(1) Extract preamble pilots by assuming a coarse carrier frequency offset.
(2) Multiply one pilot by the complex conjugate of the next pilot to demodulate the differential modulation.
(3) Take a correlation of the expected differential demodulation result with the calculated results from the received data.
(4) Repeat steps 1 to 4 for all carrier offsets within the search range (using a subcarrier step size).
(5) Search for the maximum correlation result which gives the coarse AFC.

This approach suffers in a frequency selective channel. However, the effect of the frequency selective channel is expected to be very small, since the expected channel impulse response delay spread is much shorter than the OFDM symbol period (less than 1/64 or 1/128), hence the correlation bandwidth is much wider than 6 subcarriers (the preamble pilot spacing).

<u>Using continual pilots</u>

Continual pilots are at pseudo-random locations that are unique to its carrier frequency within the 7.61 MHz range. We can use the pseudo-random locations for the coarse AFC as long as the carrier frequency offset is within half of the 7.61 MHz range, which is enough range for most practical cases. The procedure for this approach would be as follows:

(1) Extract two sets of continual pilots from two consecutive OFDM symbols by assuming a coarse carrier frequency offset.
(2) Multiply one set by the complex conjugate of the other set, taking the summation of the multiplication results.
(3) Repeat steps 1 and 2 for all carrier offsets within the search range (using a subcarrier step size).
(4) Search for the maximum result which gives the coarse AFC.

This works because only continual pilots are expected to have the same data from one symbol to the next symbol at the same carrier location. Additionally, the continual pilot locations are designed to have a good auto-correlation characteristic.

Unlike the preamble pilot approach, using continual pilots doesn’t suffer from a frequency selective channel, however, it does suffer from sampling rate offset instead. This is because this approach takes the difference (differential decoding) between OFDM symbols followed by a summation across frequency (typically 7.61MHz). The differentially decoded result could be phase rotated across frequency due to a sampling frequency offset. If the amount of phase rotation reaches 360 degrees then the summation results would be zero, since each correlation result pointed to all angles equally. This can be avoided without complex processing. Simple approaches include limiting the sampling frequency range or splitting the summation block into smaller frequency ranges.

9.3.2 Fine AFC / CPE estimation / sampling frequency offset

We are going to treat Fine AFC, CPE Estimation and Sampling Frequency Offset together since the detection of these is based on a phase rotation from one symbol to the next symbol, which can be estimated using the continual pilots.

The common process for all of these estimations is that of extracting the continual pilots and measuring the phase rotations from the previous symbol. The following figure shows one example of these phase rotations, with frequency as on the x-axis.

From the figure, we can see two kinds of information. One is 'Common Phase Rotation', which is the average phase rotation from one symbol to the next symbol (i.e. the y-intercept). The other is 'Phase Rotation Slope', which is the slope of the phase rotation across frequency. The 'Common Phase Rotation' is used for Fine AFC and CPE Estimation, while the 'Phase Rotation Slope' is used for Sampling Frequency Offset estimation. Further details on each of these are presented in the following sections.

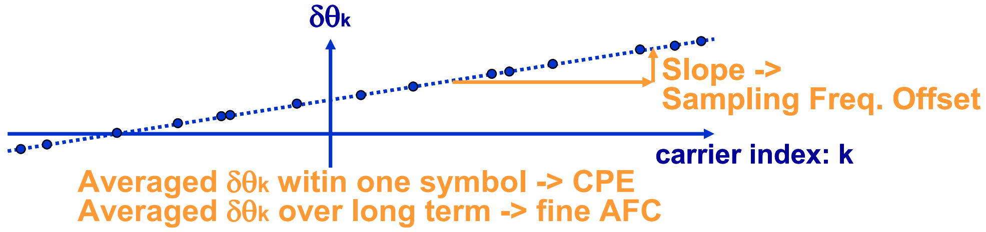


Figure 66: Effect of phase rotations per sub-carrier index k

9.3.3 Fine AFC / CPE estimation

Fine AFC is Automatic Frequency Control within one subcarrier spacing, while CPE estimation is Common Phase Error estimation, which estimates the low frequency component of a phase noise and can be cancelled subsequently based on the estimation. Both of these are estimated based on a common phase rotation across frequency.

Estimation of the common phase rotation is simply performed using the continual pilots, which are inserted continually in time, hence we can easily estimate the phase rotation from one symbol to the next for a given continual pilot carrier location. Then averaging over all the phase rotations within one symbol gives us the common phase rotation.

The phase rotation due to fine carrier offset is constant over many symbols, hence we can use long term averaging to extract accurate fine carrier offset. On the other hand, CPE contains instantaneous changes of the phase rotation, hence we cannot use long term averaging in the time direction, but we can still average within one OFDM symbol.

9.3.4 Sampling frequency offset

Sampling Frequency Offset estimation is based upon the fact that sampling frequency offset causes symbol timing drift from one symbol to the next. The effect of symbol timing drift in the frequency domain is a phase rotation slope across frequency.

The process of estimating the phase rotation slope is also performed using the continual pilots, and once the phase rotations across frequency are estimated, the slope of the phase

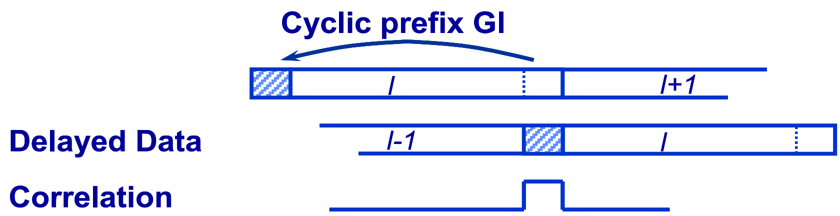


Figure 67: Utilization of Guard Interval correlation for symbol timing recovery

rotation can be derived. Because the phase rotation slope due to a sampling frequency offset is constant over multiple symbols, we can also use long term averaging to extract an accurate symbol frequency offset.

9.3.5 Symbol timing

OFDM receivers need symbol timing alignment in order to perform the FFT on appropriate data. This is relatively straightforward if there is no multipath or very short multipath compared to the GI length, since the DVB-C2 cyclic prefix (GI) provides margin for the symbol timing alignment. In the case of long multipath conditions it becomes a more difficult task, since the margin is reduced and in the case that the echo delay is equal to or greater than the GI length there could be no margin at all.

Two approaches are going to be introduced in the following sections. One is using GI correlation, which doesn't actually involve using pilots at all, but it is a very common approach for symbol timing detection so will be discussed first. The second approach is using the time domain channel impulse response (CIR), which in turn relies on the FFT output data, hence it requires reasonably good symbol timing to start with. The merit of this approach is that it gives a very accurate picture of the symbol timing.

Guard Interval (GI) Correlation

Because DVB-C2 has a cyclic prefix (GI), the end of the symbol is same as the start of the symbol. Therefore, if we take a correlation between the received data and the delayed received data, the correlation gives a strong peak at the end of the symbol. This peak can be used to calculate symbol timing.

GI correlation works on time domain data, hence it does not require computing the FFT. This is a very strong and useful technique to detect symbol timing. Under some channel conditions, however, it doesn't give a very accurate symbol timing estimate, hence we need another technique to improve the detected symbol timing. Time domain CIR is one such technique used to improve the symbol timing. This will be explained in the next section.

Time Domain CIR Estimation

This technique relies on the FFT output data to detect symbol timing, which means we need another technique to acquire an initial symbol timing estimate in order to start FFT processing (e.g. GI correlation). Once we obtain an initial symbol timing estimate and start FFT processing, this technique gives very accurate symbol timing information.

Previous sections discussed deriving a channel estimate. Although primarily used for equalisation, the channel estimate can also be used for symbol timing. The time domain CIR is obtained by applying an IFFT to the frequency domain channel estimation result. The result shows the echo locations in time, with the origin representing the current symbol timing. From this time domain CIR data we can derive how much the symbol timing should be shifted from its current position.

This technique gives quite an accurate picture of the current symbol timing alignment, but there is an issue we need to be aware of. Because the channel estimation in the frequency domain is based upon sampled channel estimations with a certain subcarrier spacing (e.g. it has Dx spacing for temporal and frequency interpolation case, and Dx*Dy spacing for frequency only interpolation case), the time domain CIR only has a limited range, the so called Nyquist limit. Within this limit, the CIR is correctly estimated. On the other hand, an echo outside of this range is aliased back into this range and confuses symbol timing. It is therefore important to keep all channel impulse responses within the Nyquist limit range.

The following table shows the Nyquist limit for the four cases (two GI periods, and two channel estimation methods). Assuming all echoes are within the GI period, we need a Nyquist limit range two times longer than the GI period to resolve all possible aliases. Unfortunately, the frequency only interpolation technique does not provide enough range relative to its GI period. Under these conditions we need some additional information from another source to resolve the aliases. One possibility would be to use the preamble pilots, which are inserted with a higher density than the scattered pilots, hence it has a much wider Nyquist limit range.

GI frac.	1/64		1/128	
GI period	7 usec		3.5 usec	
EQ Type	Time & Freq.	Freq. Only	Time & Freq.	Freq. Only
Sampling spacing	Dx = 12	Dx Dy = 48	Dx = 24	Dx Dy = 96
Nyquest limit range	Tu/Dx = 37.33 usec	Tu/DxDy = 9.33 usec	Tu/Dx = 18.67 usec	Tu/DxDy = 4.67 usec

8MHz Channel Raster Parameter
(Useful symbol duration Tu = 448 usec)

Figure 69: Nyquist limits of echo delays

9.4 Usage of Preamble Pilots

In the previous section we mainly dealt with the usage of pilots in the data symbols, only

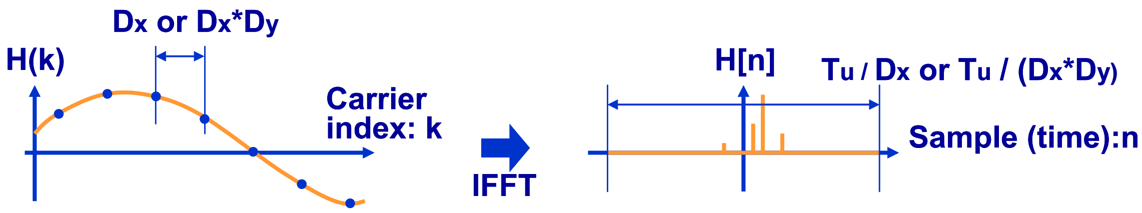


Figure 68: CIR estimation in frequency domain (left) and time domain (right)

briefly touching on the usage of the preamble pilots. In this section we present a few points related to the usage of preamble pilots.

Coarse AFC / Frame Synchronisation

In the previous section we discussed coarse AFC using preamble pilots, which uses differential decoding of the preamble pilots. Exactly the same technique can be applied to detect the frame boundary, i.e. for frame synchronisation.

Channel Estimation

Since the preamble pilots are inserted continually in time, no temporal interpolation is required. Only frequency interpolation is required in order to perform channel estimation.

Fine AFC / CPE Estimation / Sampling Frequency Estimation

Again, since the preamble pilots are inserted continually in time, we can use the same techniques as used on data symbols which utilise the continual pilots.

Symbol Timing

The time domain CIR approach to symbol timing works well with the preamble pilots since they are inserted with much higher density, hence it has a very wide range based on the Nyquist limit.

9.5 Notes for pilots

This section provides some additional notes regarding the pilots.

Pilot Boosting

- CP / SP / EP all have the same boosting factor of 7/3
- Preamble pilots have a different boosting factor to minimise the averaged symbol power differences

Pilot Modulation

- BPSK modulation
- Modulation phases are dependent upon the subcarrier index and PRBS sequence
- The sequence starts from zero subcarrier frequency
- All pilots share the same PRBS sequence

CP Location

- Pseudo-random location repeated every 7.61 MHz
- The 7.61 MHz repetition block is aligned with its preamble block
- All continual pilots are bearing preamble pilots, means continual pilot carriers are always preamble pilots in preamble symbols. Hence some of synchronisations that are based on the continual pilots (e.g. fine AFC and sample rate offset estimations) can be carried out without the frame synchronisation.

10 Concept of Framing

As far as reasonable many parts of the DVB-C2 standard are adopted from other DVB 2nd generation physical layer standards (DVB-S2,-T2). Nevertheless especially the framing concept contains some special properties that reflect the specific requirements of the cable environment. For understanding, many properties of the C2 framing it is useful to start from the functional requirements at the beginning of the C2 standardization process: First of all and most important there was a desire for a significantly increased spectrum efficiency. The C2 standard should be designed to minimize all overhead that is needed for topics like adjacent channel isolation, channel estimation, inter-symbol interference avoidance and many more. Furthermore an increased flexibility has been requested to handle different input streams with very different input bandwidths without stuffing overhead. There was a high level of consensus from the beginning that similar to DVB-T2 a preamble to signal physical layer parameters of the C2 channel should be included. In that preamble all C2 channel parameters must be accessible in any operating condition, just to mention operation states as initial switch-on or different tuning positions. Finally, an efficient notching mechanism that allows omitting C2 spectrum frequencies that overlap to critical terrestrial services (i.e. aviation radio services) has been requested. The mapping of the upper requirements into the C2 framing is explained in this chapter.

10.1 Broadband transmission signals and segmented OFDM reception

One reason for the increased spectrum efficiency of the OFDM based C2 system is – beside other factors like guard interval size or pilot density – the relative size of the Guard Band that is needed to isolate adjacent signals from the C2 signal (and vice versa). DVB-C2 is based on the 4k OFDM mode from DVB-T2, adopting the subcarrier spacing of (64/7 MHz)/4096 = 2,232 kHz. It is one of the 'magic' formulas of OFDM that the useful OFDM symbol duration is the inverse of the subcarrier spacing. For DVB-C2 this is 1/2,232 kHz = 448 µsec. It is important to notice that the useful OFDM symbol duration remains constant independent from the actual number of OFDM subcarriers as long as the carrier spacing is kept constant. On the other hand the OFDM spectrum shape stays almost constant independent from the number of used OFDM subcarriers. Therefore it is obvious that the larger the C2 transmission signal (i.e. the more used OFDM subcarriers) the lower the relative Guard Band overhead. As a result it is beneficial to use broader transmit spectrums.

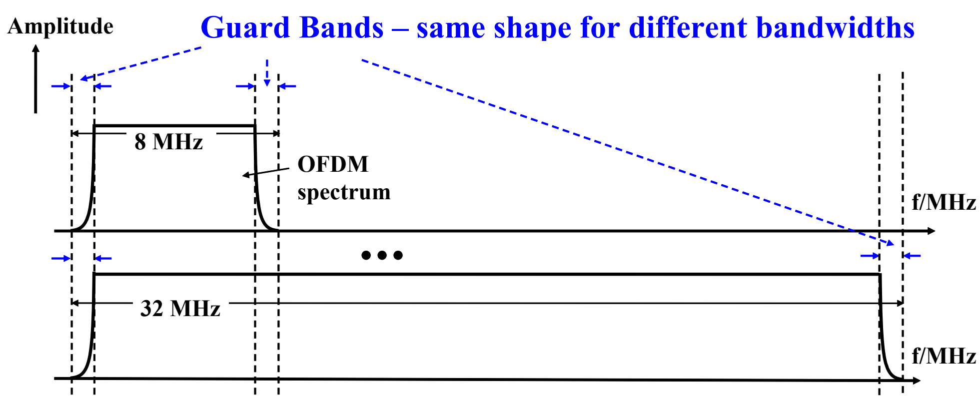


Figure 70: Effect of lower Guard Band overhead with increased OFDM spectrum bandwidths

As an example Figure 70 shows schematically the C2 signal spectrum for 2 different signal bandwidths (8 MHz and 24 MHz). The shape at the signal border remains the same for both bandwidths.

The advantage of broader OFDM transmit spectrums has been motivated in the previous section. While the larger signal generation complexity of an increased bandwidth (i.e. larger IFFT size) is acceptable for the cable headend side, the complexity has to be limited for the consumer device on receiving side. This becomes possible in DVB-C2 with the principle of segmented OFDM: The receiver selects a certain bandwidth area out of an overall broader transmit spectrum that includes the service it wants to decode. Segmented OFDM takes advantage of the constant useful OFDM symbol duration. Figure 71 illustrates the mechanism.

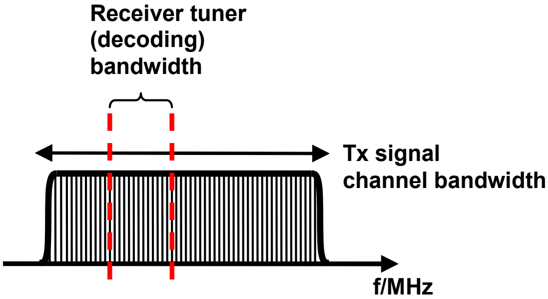


Figure 71: Segmented OFDM principle - the receiver extracts a part of an overall broader transmit signal

The resulting advantage is that although the C2 transmit signal typically will be broader than the reception window a standard consumer electronic tuner can be used on reception side. DVB-C2 is designed to allow a basic reception scenario with a single 8MHz tuner. However, broadband reception scenarios beyond 8MHz are possible (Data Slice bundling) and will be discussed in the following subchapter.

10.2 Framing structure

Figure 72 shows the DVB-C2 framing structure in time and frequency domain. Each C2 frame begins with one or more preamble symbols, followed by a fixed number of 448 data symbols. Depending on the number of preamble symbols and the chosen Guard Interval the overall C2 frame duration slightly changes around 200msec. The preamble symbols are divided in frequency direction into 7.61 MHz L1 blocks (i.e. equivalent to the bandwidth of 3,408 subcarriers). The embedded L1 information is spread over the subcarriers of L1 blocks. Each L1 block contains the same signalling information.

The underlying Data Slices that follow the preamble symbols have a flexible bandwidth and follow no fixed frequency raster. The bandwidth of the Data Slices only depends on a Guard Interval specific granularity (i.e. 12 subcarriers for GI=1/64 or 24 subcarriers for GI = 1/128). Data slices must not exceed a bandwidth of 7.61 MHz in order to allow segmented OFDM reception with a standard 8 MHz CE tuner. The bandwidth of a C2 signal is at least 7.61MHz to cover one complete L1 block and can go up to bandwidths of roughly 450 MHz.

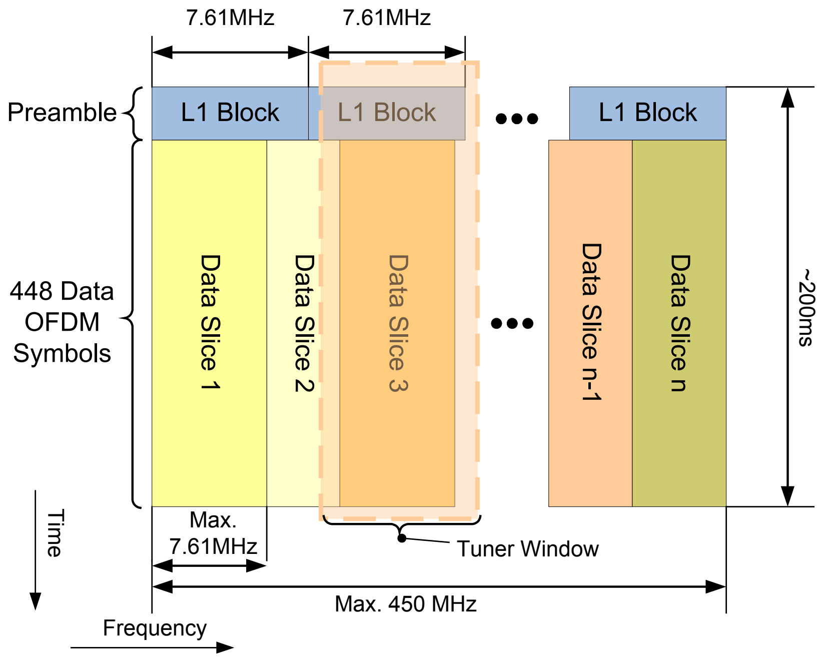


Figure 72: DVB-C2 framing structure

Typically all XFECFRAME packets of a PLP (physical layer pipe) are located in a single Data Slice, which can contain data of one or several PLPs. Figure 72 shows exemplary the tuning position of a receiver that wants to decode XFECFRAME packets of a PLP that is inserted in Data Slice 3. The typical Data Slice specific decoding procedure on receiver side is as follows: In a first step a reception FFT is applied (e.g. 4k FFT) over the complete tuning window. Afterwards the receiver selects and decodes only these carriers that belong to the desired Data Slice.

Another feature of the C2 standard is the so called PLP bundling. It is intended for non broadcast applications and target high bit rate services such as broadband internet: In this case PLPs with very high data rates are distributed over several Data Slices. Theoretically a single PLP might allocate the overall C2 signal bandwidth, but of course the mixture with other PLPs in the same and other Data Slices remains possible. If PLP bundling is applied the receiver requires either several 8MHz tuners or alternatively tuner architectures with a broader reception window. The XFECFRAME packets of the bundled PLP can be reordered from the different Data Slices with the help of a counter mechanism (ISSY counter).

As a conclusion the C2 standard is able to handle a varying number of input streams of very different bandwidths without stuffing overhead – starting from narrow single Data Slices up to a large number of bundled Data Slices.

10.3 Decoding of the Preamble Symbol Data

As explained in the previous chapter Data Slices can have an almost arbitrary bandwidth while L1 blocks are aligned to a fixed 7.61 MHz raster. The reason for the separate bandwidth allocation is the need to access all signalling information in all operating conditions, including any tuning position inside the C2 spectrum as well as the initial acquisition after switching on the receiver. The L1 blocks of the preamble carry all Layer 1 information required to decode the payload data in the Data Slices. The information inside the L1 blocks is cyclically repeated every 3408 subcarriers. If the receiver's tuner window is not aligned to the L1 Block structure in the frequency domain, the receiver is able to obtain the data by sorting the data of two blocks. As a result the complete signalling is available in any tuning position within the C2 signal. This is an important feature that allows supporting Data Slices with very different bandwidths at very different frequencies. Figure 73 shows the basic procedure: The tuning position to decode a specific Data Slice includes parts of two neighboured L1 Blocks m and m-1. The L1 information is cyclically repetitive modulated on the subcarriers of the different L1 blocks. As a result, the complete L1 signalling can be retrieved by applying the reception FFT, taking the parts of the two L1 blocks and reorder them accordingly. At the output of the reordering stage the complete L1 block is available and can be decoded and evaluated.

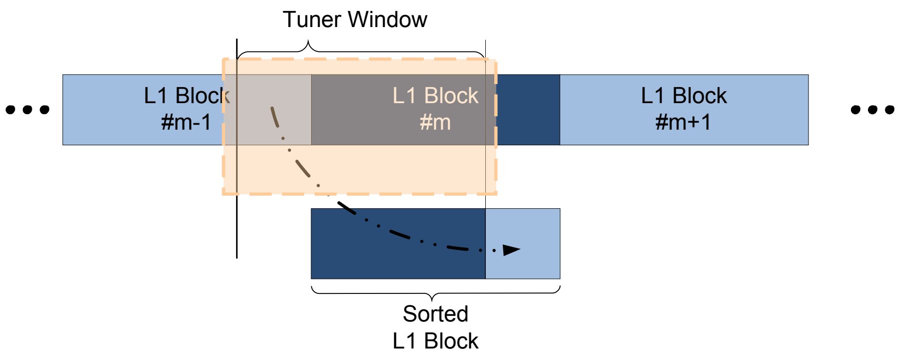


Figure 73: L1 block reordering to retrieve L1 signalling (after reception FFT)

Another feature of the increased spectral efficiency of the C2 standard is the possibility to vary the Data Slice bandwidth within the tuning window from C2 frame to C2 frame. This becomes possible since the actual tuning position for receiving a Data Slice is signalled in the L1 information. Therefore all receivers that are decoding this Data Slice have the same tuning position. As a result, changing the Data Slice width and/or the position within the tuning window from C2 frame to C2 frame requires no retuning and generates no service interruption. Figure 74 shows schematically a Data Slice and the overlaid tuning window, which is equal to the 'breathing' range of the Data Slice.

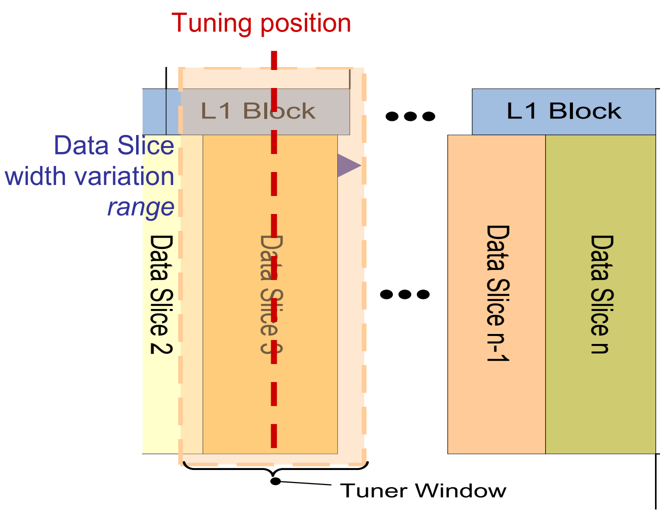


Figure 74: Varying Data Slices within the reception tuner window

10.4 Absolute OFDM concept

The concept of 'absolute OFDM' is unique to DVB-C2. The L1 Signalling Blocks begin at the absolute frequency of 0 MHz and are partitioned in steps of 7.61 MHz. In contrast to other DVB standards it is not possible to shift a C2 baseband signal to any RF carrier frequency rather than being defined in a unique way for the whole cable spectrum: Especially the pilot sequences of the OFDM signal are different for all different frequencies. The reason for that behaviour is to avoid unwanted repetitions in the frequency domain which may cause unwanted high peak values of the OFDM signal in time domain. Furthermore the unambiguous pilot sequences allow for easy and reliable synchronization and offset compensation. Although the L1 block partitioning and the related pilot sequences are defined for the whole cable spectrum – of course L1 blocks are only transmitted in these frequencies where Data Slices are present.

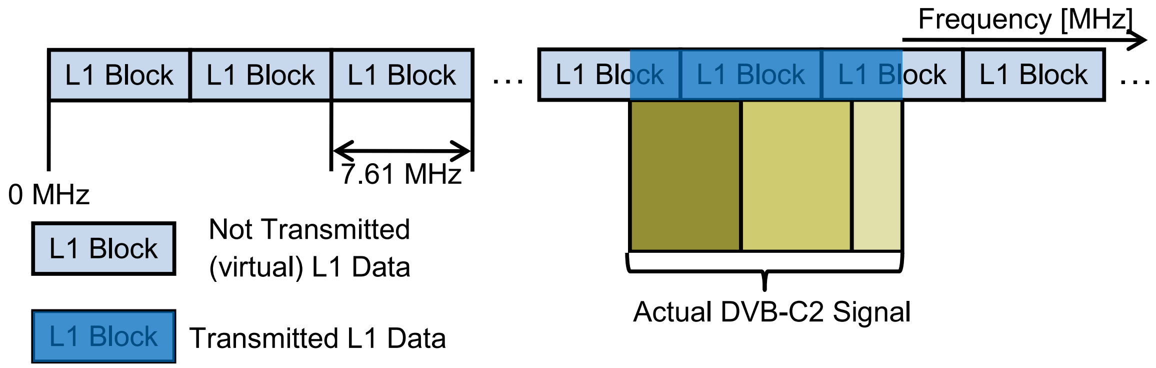


Figure 75: Absolute OFDM concept in DVB-C2

Terrestrial services and HFC cable networks often share the same frequency range (see Figure 76). Due to insufficient cable shielding, mainly in the close range of the consumer electronic device, accumulated radiation from cable networks disturbs the operation of the terrestrial services. This is especially critical for all kind of security related terrestrial services (e.g. flight security radio). Vice versa ingress from terrestrial services degrades the signal quality of services in the cable network. It is one major advantage of OFDM that only these subcarriers can be notched that actually overlap with the frequencies of the critical terrestrial services. The spectral efficiency advantage of DVB-C2 is obvious when comparing this solution to existing DVB-C systems, where sometimes complete 8 MHz channels have to be released.

10.5 Notching concepts

C2 offers two types of notches: On the one hand there are the so called 'Narrowband notches', which have a bandwidth smaller or equal than 48 subcarriers. It is mandated that only one notch per 7.61 MHz is allowed. Narrowband notches may also be located inside Data Slices. The other possibility of omitting OFDM subcarriers are the 'Broadband notches', which exceed a bandwidth of 48 subcarriers: Broadband notches are always located between Data Slices, at least at one side of the Broadband notch a consistent L1 Block is needed. Notching is applied in any case for preamble and data symbols.

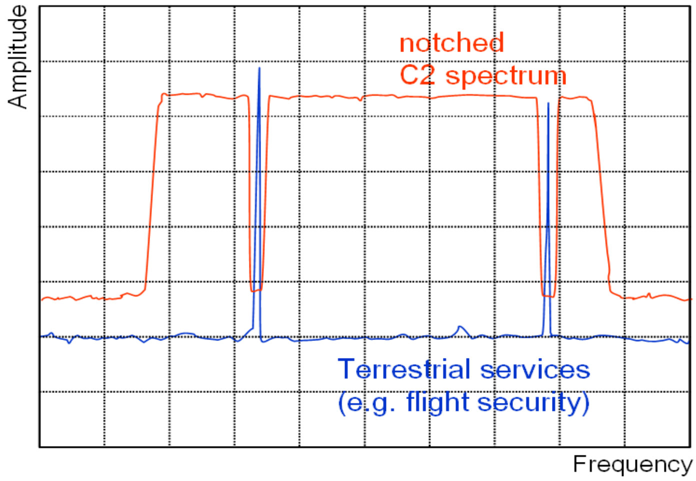


Figure 76: Notched DVB-C2 spectrum with overlapping terrestrial services

11 L1 & L2 Signalling and Preamble Structure

The L1 signalling scheme of the DVB-C2 system is supporting a flexible and efficient way by using Preamble header, L1 signalling part2 and FECFrame header as shown in Figure 77. Preamble header and L1 signalling part 2 are conveyed by Preamble OFDM symbols which are existing on the beginning of every C2 Frame and FECFrame header is existing on the beginning of every Data slice packet of the type 2 Data Slice.

Preamble header introduces the way to decode L1 signalling Part 2 and L1 signalling Part 2 has information of the configuration of current C2 Frame, Data Slices, PLPs and Notch bands. For type 1 Data Slice, all information to decode the target PLP is introduced by L1 signalling Part 2 only and it makes overhead for signalling smaller. For type 2 Data Slice, the configuration of every Data Slice Packet can vary with its modulation, code rate, FEC type etc. So FECFrame header supports the way to find the target Data Slice packet and to decode it and it gives the very flexible structure.

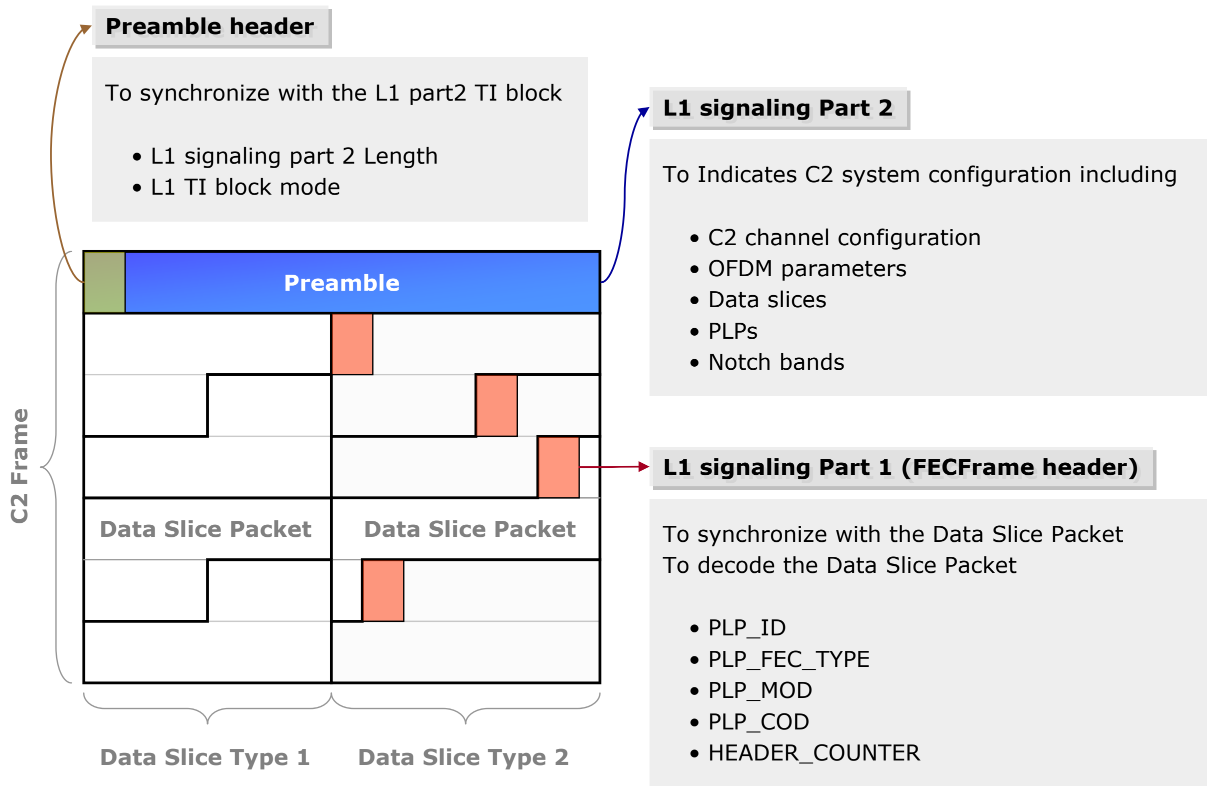
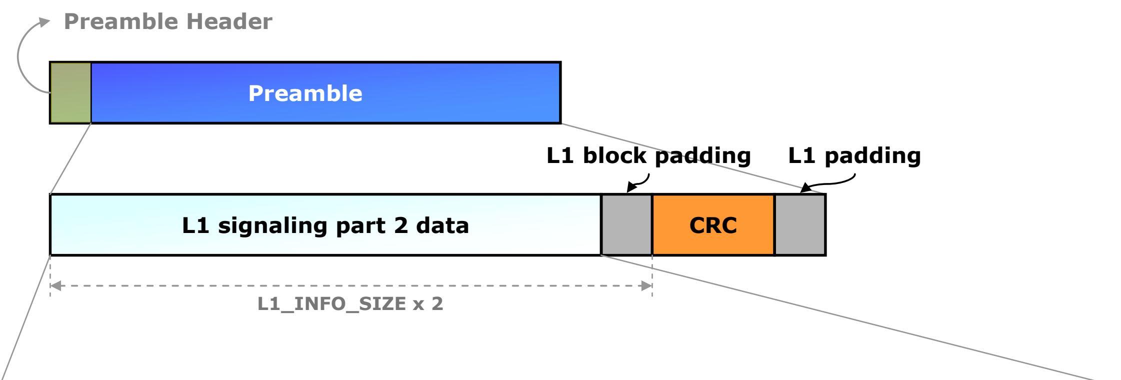


Figure 77: L1 Signalling structure

Preamble Header conveys the L1_INFO_SIZE and L1_TI_MODE. Using these 2 parameters, the size and shape of L1 TI block can be calculated and also the size of encoded L1 block can be calculated. These 2 parameters are encoded and modulated with same way as FECFrame Header process. Reed-Muller (32,16) is used for encoding and QPSK is used for its modulation. As a result, Preamble Header of 32 OFDM cells are inserted in front of L1 TI block on every Preamble OFDM symbols. In receiver side, the position of L1 TI block can be found by detecting Preamble Header. L1 Signalling Part2 consists of C2 system parameters, Data Slice parameter loop, PLP parameter loop and Notch Band parameter loop as shown in

Figure 78. The C2 system parameter part is 110 bits and identifies C2 system network, OFDM characteristics, Channel configuration etc. Also it identifies the size of Data Slice parameter loop and Notch Band parameter loop. The Data Slice parameter loop part has variable length depending on the number of Data Slice within the current C2 system, Guard interval (GI) mode and Data Slice type. In this loop, Data Slice identifier, the position and size of Data Slice, TI mode and other configuration related parameters are identified. And the number of PLPs within current Data Slice is also signalled which is defining the size of PLP loop. The PLP parameter loop part also has variable length depending on the number of PLP, PLP type, Data Slice type and PSI/SI reprocessing. This loop identifies PLP identifier, PLP bundling, Payload type, PLP type etc. In case of Data Slice type 1, PLP loop has information of only 1 PLP and conveys more detailed PLP information such as the start position of the first complete FECFrame, FEC type, modulation and code rate. So this PLP doesn't require the FECFrame header and can eliminate the overhead for it. Also the transport stream identifier and original network identifier can be conveyed in this loop if necessary. It makes easy to find relevant L2 information of current service when service from other network is retransmitted through DVB-C2 system without PSI/SI reprocessing. The Notch Band parameter loop part conveys the position and size of Notch Bands and its length is also varying depending on the number of Notch Bands and GI mode.



C2 System parameters	110 bits (define the number of Data Slice and Notch Band)
Data Slice parameter loop	Variable depends on number of Data Slice, GI mode, Data slice type
PLP parameter loop	Variable depends on number of PLP, PLP type, Data slice type, PSI/SI reprocessing
Notch Band parameter loop	Variable depends on number of Notch Band, GI mode

Figure 78: L1 signalling Part2 structure

L1 block padding can be inserted following L1 signalling part 2 data, if necessary, to ensure that the position of CRC can be identified with L1_INFO_SIZE. The L1_INFO_SIZE is 14bits but it needs at least 15bits to signal the maximum size of L1 signalling part 2 data. To solve this problem, the length of L1 signalling part 2 data is identified with double of L1_INFO_SIZE. Consequently, when actual size of L1 signalling part 2 data is not multiple of 2, it needs 1 bit padding to guarantee the position of CRC is identified with L1_INFO_SIZE. And Maximum length of L1 signalling part 2 data is limited to 32766bits. At the end of CRC, L1 padding can be inserted, if necessary, to ensure that multiple LDPC blocks for L1 part2 have same size.

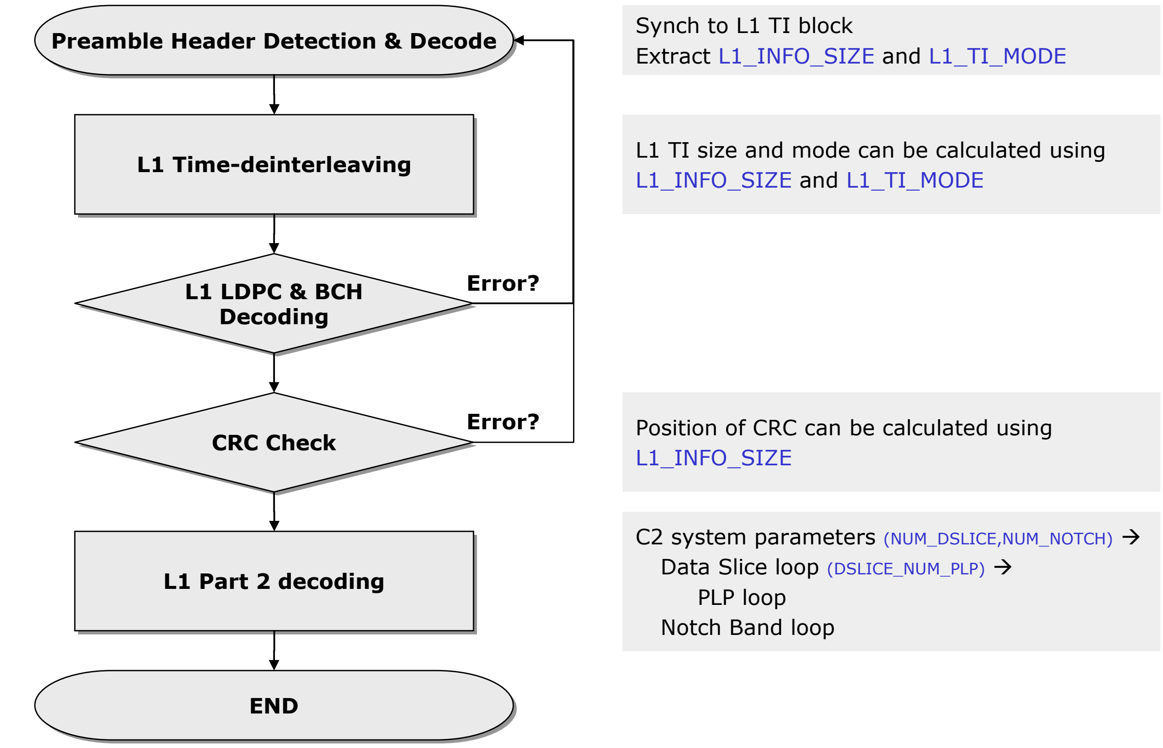


Figure 80: L1 signalling Part 2 decoding sequence

The L1 signalling Part 2 decoding sequence is shown in Figure 80. To decode L1 signalling Part 2, Preamble Header should be detected and decoded at first. By doing this, receiver can synchronize to L1 TI block and can do L1 time de-interleaving using L1_INFO_SIZE and L1_TI_MODE. After that error bits can be corrected and checked by LDPC/BCH decoding and CRC check. To get parameters of the punctured and shortened LDPC block and to get CRC position, L1_INFO_SIZE can be used. And receiver gets L1 signalling Part 2 data without error bits.

Syntax	No. of bits	Identifier
c2_delivery_system_descriptor {		CORE
descriptor_tag	8	uimsbf
descriptor_length	8	uimsbf
descriptor_tag_extension	8	uimsbf
plp_id	8	uimsbf
c2_system_id	16	uimsbf
if (descriptor_length > 4) {	Transmission & Frequency	
C2_System_tuning_frequency	32	bslbf
active_OFDM_symbol_duration	3	bslbf
guard_interval	3	bslbf
reserved	2	bslbf
}		
}		

Figure 81: C2 delivery system descriptor

DVB-C2 system has C2 DSD (Delivery System Descriptor) which is in line with the other DVB delivery system descriptors as shown in Figure 81. C2 DSD consists of the core part and the extension part. The core part is describing the C2 system and PLP identifier while the extension part is describing the transmission mode and tuning frequency. So C2 DSD maps the streams to the relevant tuning frequency for the targeted Data Slice and PLP.

Figure 82 shows the tuning process from NIT information. At first, receiver should be tuned to targeted C2_System_tuning_frequency using C2 DSD and get DVB-C2 Frame at tuned frequency. By decoding L1 part2, receiver can check the IDs, such as network ID, C2 system ID, Data Slice ID, PLP ID, to check the current C2 system is relevant to targeted service. And receiver can re-tune to targeted Data Slice conveying relevant PLP using DSLICE_TUNE_POS of L1 signalling part 2 when it is required. The targeted PLP can be found using PLP_ID and if the type of targeted PLP is the grouped PLP, common PLP also should be found to get L2 information. Then receiver can start to decode the service.

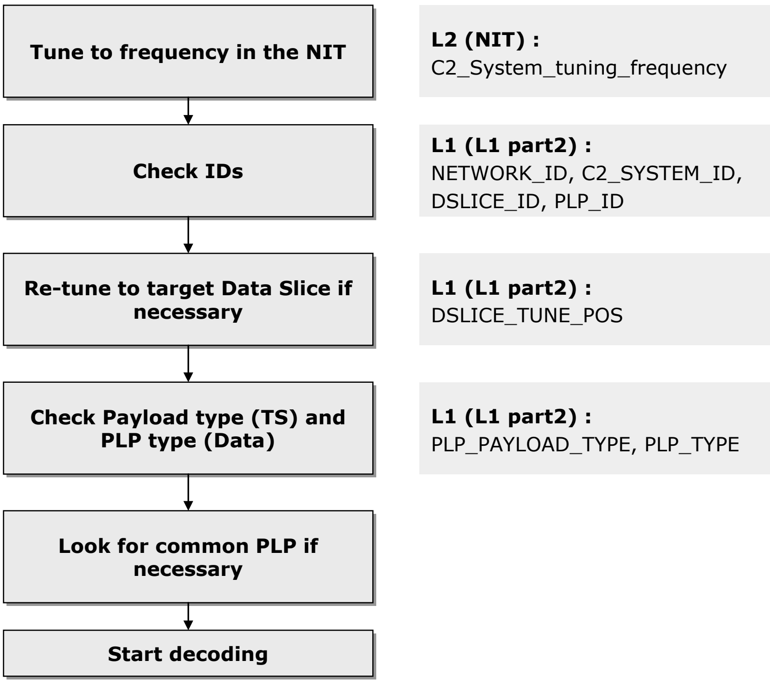


Figure 82: Tuning process from NIT

12 Preamble Signal Protection Mechanisms

The preamble signal of DVB-C2 systems (consisting of one or more OFDM symbols) carries the L1 (layer-1) signalling required for the decoding of the Data Slices and their payload, and therefore, it should be robustly protected. As depicted in Figure 83, the preamble signal consists of a frequency cyclic repetition of the L1 blocks that are repeated every 7.61 MHz and each L1 block consists of preamble header, coded L1 signalling and their copies. The repetition of the L1 blocks enables to access the complete L1 signalling in any tuning position of an 8 MHz receiver tuner. On the other hand, the repeated structure of the coded L1 signalling in each L1 block increases the robustness and reliability of the L1 signalling. Moreover, since the coded L1 signalling is protected with the coding rate and modulation order lower than those of data payloads, it is much more robust than data payloads

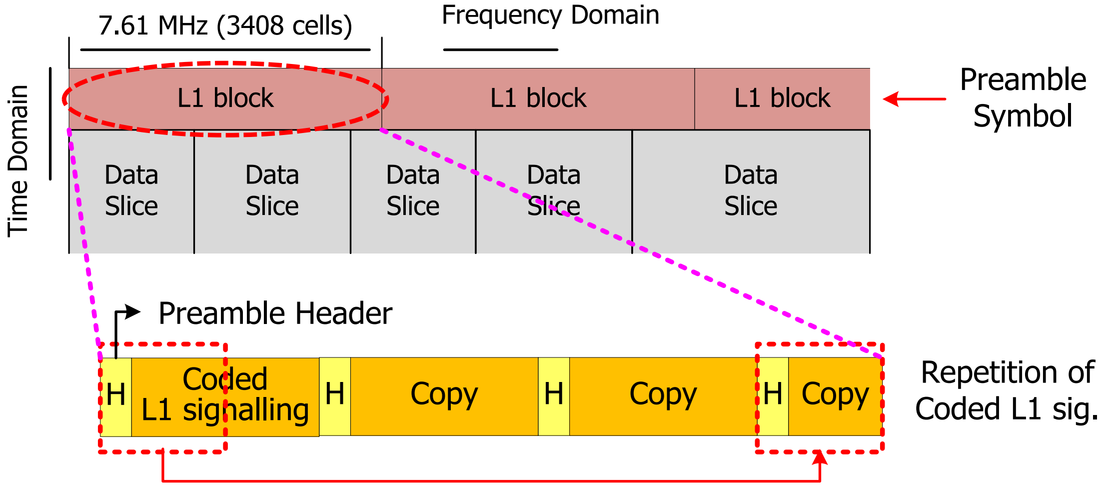


Figure 83: Structure of DVB-C2 Preamble

As shown in Figure 84, the processing of coding and modulation for L1 signalling has three additional steps, Segmentation, Zero-Padding, and Puncturing, as compared with the protection of data payloads. The size of the L1 signalling bits in DVB-C2 systems varies depending on the complexity of the underlying Data Slices and it can be up to approximately tens of thousands. If it exceeds a predetermined number, a segmentation operation is applied since the number of available OFDM carriers is not large enough to transmit such a big L1 signalling data through one OFDM symbol.

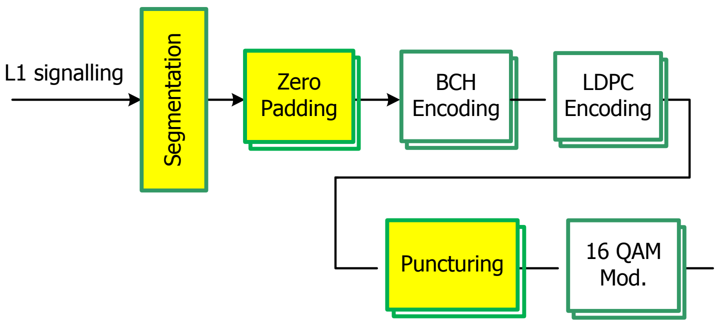


Figure 84: Block-diagram for preamble protection

Each segmented L1 signalling is protected by a concatenation of BCH outer and LDPC inner codes. However, since the size of each segmented L1 signalling is always less than the information size for the BCH code, a zero-padding (or shortening) of information bits is applied for BCH encoding. Furthermore, a puncturing of LDPC parity bits is applied for controlling the robustness of the coded L1 signalling.

12.1 Segmentation

In DVB-C2 systems, if the size of L1 signalling exceeds a predetermined number, the L1 signalling is divided into multiple equidistant blocks, as depicted in Figure 85. Furthermore, each segmented block is protected by a concatenation of BCH outer and LDPC inner codes and the coded bits are mapped into 16 QAM symbols. After coding and modulation, the preamble header is appended to each block and finally, the L1 block and each preamble symbol are determined as shown in Figure 85.

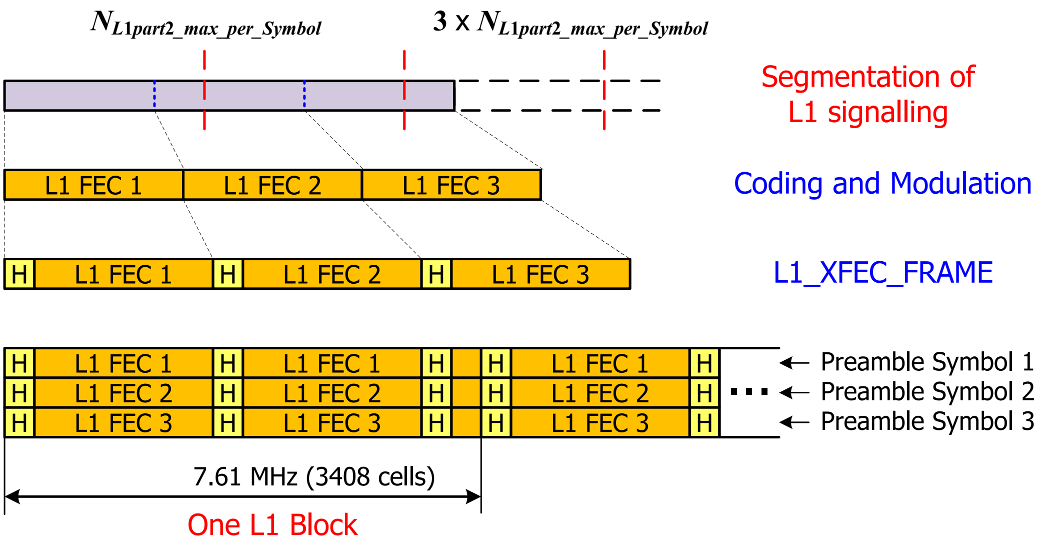


Figure 85: Segmentation of L1 signalling and construction of preamble symbols

12.2 Zero-padding and puncturing

The zero-padding of BCH (or LDPC) information bits and the puncturing of LDPC parity bits are performed bit-group by bit-group in the predetermined order. In other words, the positions of the zero-padding and puncturing bits for BCH and LDPC encodings are allocated according to the bit-groups. The bit-group is differently defined for information and parity bits, respectively, as shown in Figure 86 and Figure 87. For the zero-padding, the BCH information bits are successively divided into 20 bit-groups which are 19 groups of 360 bits and 1 group of 192 bits. For the puncturing, the LDPC parity bits are periodically divided into 25 groups of 360 bits. Here, the grouping factor 360 and the grouping patterns are derived from the algebraic characteristic of DVB-S2/T2/C2 LDPC codes (so-called '360-periodicity'). DVB-C2 adopts the group-based zero-padding and puncturing by theoretical analysis and computer simulations. Consequently, the DVB-C2 system can provide a stable and good error performance for the coded L1 signalling.

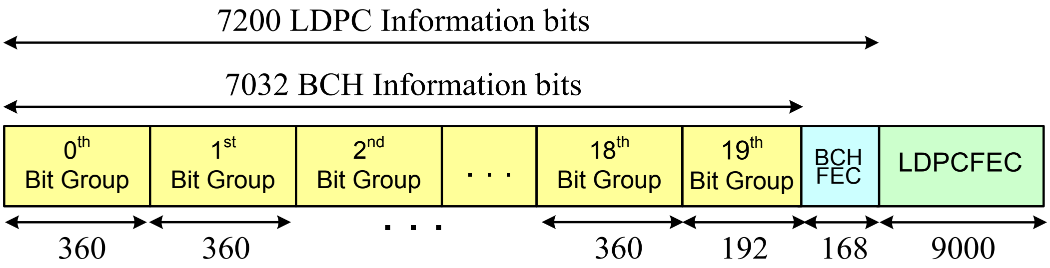


Figure 86: Bit-groups of BCH (or LDPC) information bits for zero-padding

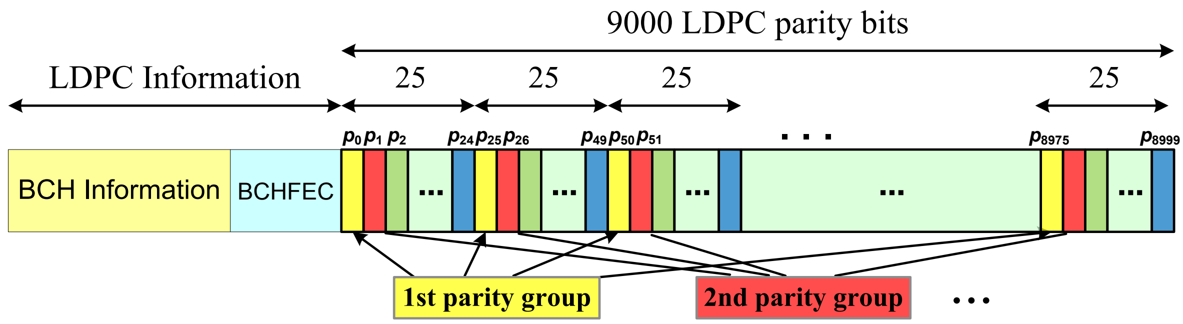


Figure 87: Bit-groups of LDPC parity bits for puncturing

12.3 Code-rate control

Another characteristic of the preamble protection in the DVB-C2 system is the code-rate control of the coded L1 signalling. In DVB-C2, the performance variation induced by the variable size of L1 signalling bits is reduced by adjusting the rate of coded L1 signalling. As shown in Figure 88, its effective LDPC code-rate tends to decrease as the size of pure L1 signalling bits decreases. In general, for a fixed code-rate, the code performance is degraded as the information size decreases since the code size also decreases. On the other hand, for a fixed information size, it is clear that the code performance is improved as the code rate decreases. From these facts, the rate of the coded L1 signalling in DVB-C2 is systematically controlled by adjusting the number of LDPC parity bits to be punctured. Consequently, the DVB-C2 system can provide approximately invariant and stable receiving coverage for the L1 signalling, regardless of its size.

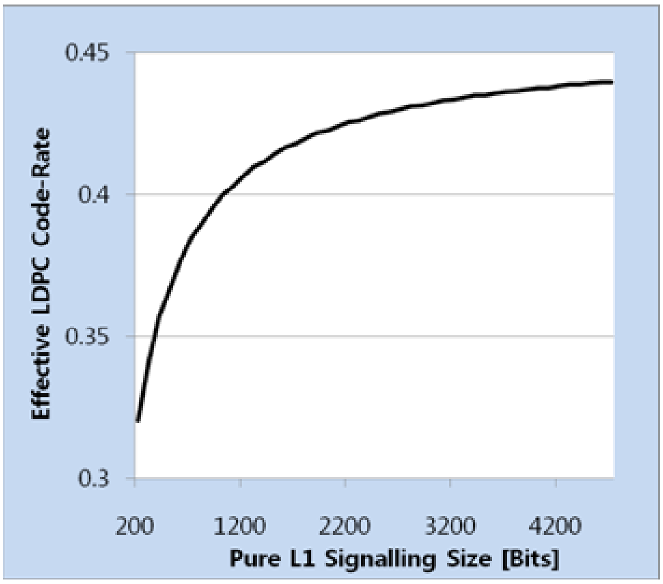


Figure 88: Effective LDPC code-rate of coded L1 signalling

13 OFDM Generation and RF Characteristics

13.1 OFDM generation

The basic block diagram for OFDM generation is shown in Figure 89.

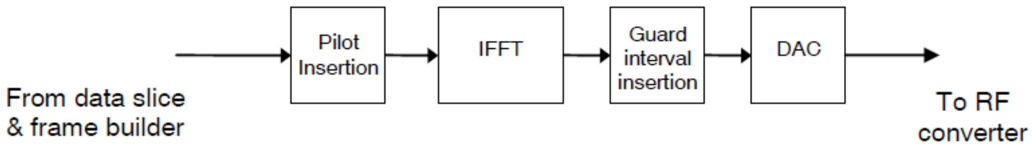


Figure 89: Block diagram for OFDM generation

The core OFDM building blocks are IFFT (Inverse Fourier Transformation) and Guard Interval insertion as already known from existing standards like DVB-T and DVB-T2.

13.1.1 Reasons for choice of OFDM for DVB-C2

In the first generation digital cable transmission standard called DVB-C, a single carrier QAM is used. Also at the beginning of the DVB-C2 technical standardisation process, the single carrier QAM technique has been considered. However, extended comparisons with OFDM resulted in several important advantages of OFDM which are outlined in the following.

13.1.1.1Advantages for transmission capacity

First advantage is the capacity gains as shown Figure 90. Single carrier QAM as used for DVB-C applies roll-off impulse shaping to avoid inter symbol interference (ISI). This kind of impulse shaping leads to Guard Bands in the frequency spectrum, which are dependent on the roll-off factor applied (15 % for DVB-C). The maximum possible symbol rate in an 8 MHz channel usable for DVB-C is 6,96 Msps (generally 6,9Msps is applied), which results in an unused spectrum of about 1MHz. Furthermore the amount of unused spectrum is dependent on the channel bandwidth. It increases if the channel bandwidth grows. The DVB commercial requirements for DVB-C2 (see chapter 2) ask for a minimum capacity gain of 30 % obtainable in existing cable networks. About 6 dB gain generated by the new LDPC FEC allows to use in existing networks a 1024-QAM with DVB-C2 instead of a 256-QAM with DVB-C without decreasing the CNR margin at the user outlet. By using single carrier QAM and an a roll-off factor of 15% only a capacity gain of 22 % compared to DVB-C could be achieved.

In contrary for OFDM no roll-off impulse shaping is necessary. For an 8 MHz channel only a Guard Band of 390 kHz must be foreseen due to spectrum characteristic of the OFDM signal (see Figure 96).

The DVB-C2 commercial requirement asking for a minimal increase in capacity of 30 % reached in existing cable networks is achieved already when applying an 8 MHz channel. In case a higher channel bandwidth is used, the capacity increases further compared to DVB-C (for example by 35 % if four 7.61 MHz Data Slices are combined within a 32 MHz wide

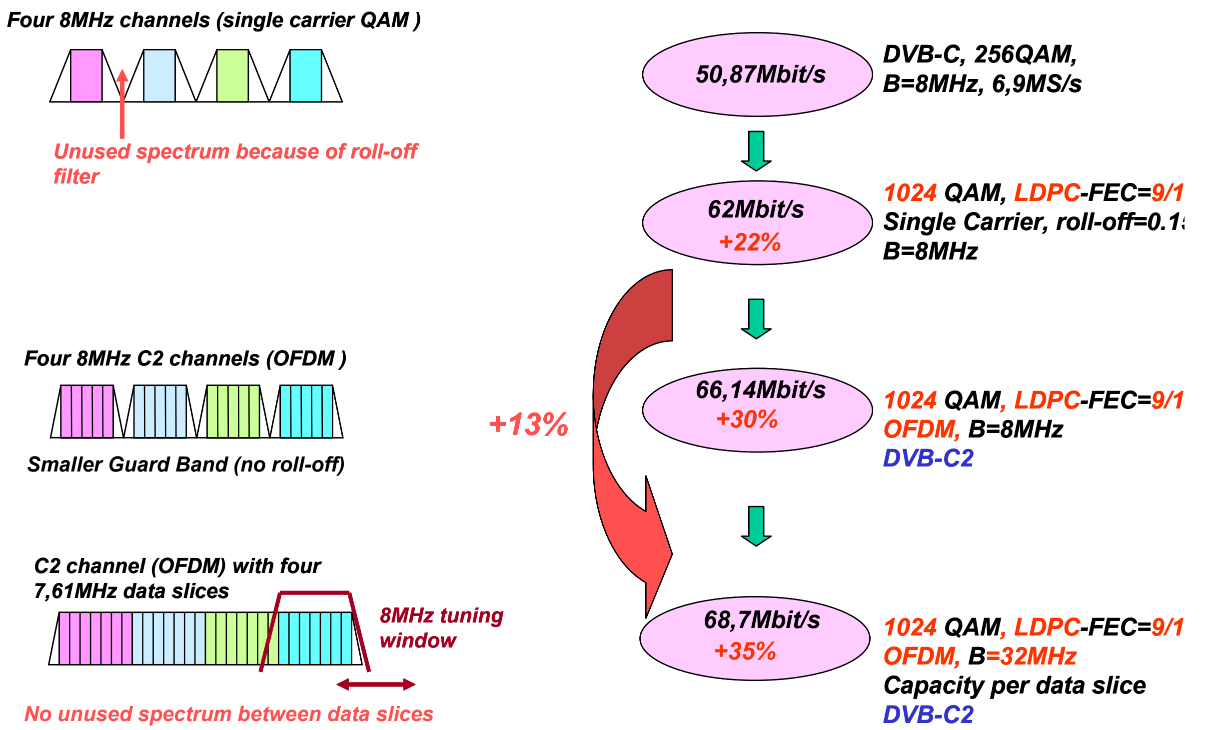


Figure 90: Capacity gain for OFDM compared to Single Carrier QAM

channel as shown in Figure 90). The reason for this phenomenon is the fact that the bandwidth of the OFDM Guard Band remains unchanged with increasing channel bandwidth. The overall additional capacity gain of DVB-C2 caused by the OFDM system is equal to 13 % compared to DVB-C's single carrier QAM. Also for reception of broader DVB-C2 channels, a set-top-box with 8 MHz tuner bandwidth can be applied, because the maximum bandwidth of the Data Slices belonging to a DVB-C2 channel is 7.61 MHz.

13.1.1.2 Flexibility in spectrum occupation

One reason why DVB-C2 can be used in a very flexible manner is the possibility to switch off OFDM carriers ("notching"). This is a helpful feature to reduce or eliminate emissions in frequency ranges of wireless security services like aircraft and civil service radio. No waste of spectrum occurs because there is no need to switch off an entire signal but only the OFDM carries concerned. This behaviour is depicted in Figure 91. Furthermore notching may be applied in the case of external ingress for example of mobile phone services.

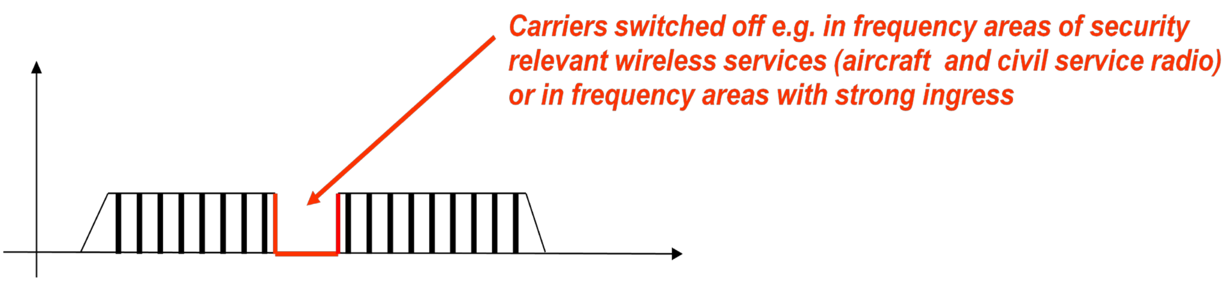


Figure 91: Flexible and efficient use of spectrum to avoid occupation of sensitive frequency areas

13.1.1.3 Some further advantages of OFDM

Further advantages of the use of an ODFM technique are:

- Simple echo cancellation due to Guard Interval (3,5 µs for GI=1/128, and 7us for GI=1/64). For most cable networks in practice, echo delays scarcely are higher than 1-2 µs.
- Robustness against narrowband interferers as they occur for example in mixed occupation scenarios with analogue TV (e.g. PAL) signals. This robustness is achieved by frequency interleaving which can be easily applied for OFDM.
- Simple equalisation in the receiver (only 1 tap equalizer necessary, no decision feedback equalizer required).

13.1.2 OFDM parameters

For DVB-C2, a carrier spacing according to 4K OFDM (within a minimum bandwidth of 6 and 8 MHz, respectively) has been chosen, which is a good compromise between realisation efforts, on the one hand, and RF characteristics (for example spurious emissions into adjacent channels) , on the other hand. Furthermore the Guard Interval values of 1/64 and 1/128 do not significantly reduce the available transmission capacity and the minimum Guard Interval time of 3,5 µs (for GI=1/128) is sufficient for reliable receiver synchronisation. Figure 92 shows the table with the OFDM parameters chosen for DVB-C2 like carrier spacing, symbol duration and Guard Interval duration.

Elementary period as a function of channel raster bandwidth

Channel Raster	"6 MHz"	"8 MHz"
Elementary period T	7/48 μs	7/64 μs

OFDM parameters

Parameter	"6 MHz" 1/64	"6 MHz" 1/128	"8 MHz" 1/64	"8 MHz" 1/128 ⇐ ***Guard Interval***
Number of OFDM carriers per L1 Block K_{L1}	3 408	3408	3408	3408
Bandwidth of L1 Signalling Block (see note 1)	*5.71 MHz*	*5.71 MHz*	*7.61 MHz*	*7.61 MHz*
Duration T_U	4096T	4096T	4096T	4096T
Duration T_U µs (see note 1)	*597.3*	*597.3*	*448*	*448*
Carrier spacing $1/T_U$ (Hz) (see note 1)	*1674*	*1674*	*2232*	*2232*
Guard Interval Duration Δ/T_u	*64T*	*32T*	*64T*	*32T*
Guard Interval Duration Δ/T_u μs (see note 1)	*9.33*	*4.66*	*7*	*3.5*
NOTE 1: Numerical values in italics are approximate values.				

Figure 92: OFDM parameters for DVB-C2

Some important additional remarks:

- For a given channel raster (6 MHz or 8 MHz) OFDM carrier spacing is constant and independent from channel bandwidth
- DVB-C2 channel bandwidth can be varied in steps of n · 12 OFDM carriers (n=1,2,3….), if GI=1/64 and in steps of n · 24 carriers, if GI=1/128
- Only the minimum bandwidth (6 MHz or 8 MHz) is in accordance with the cannel raster, for higher bandwidth deviation from the channel raster is possible

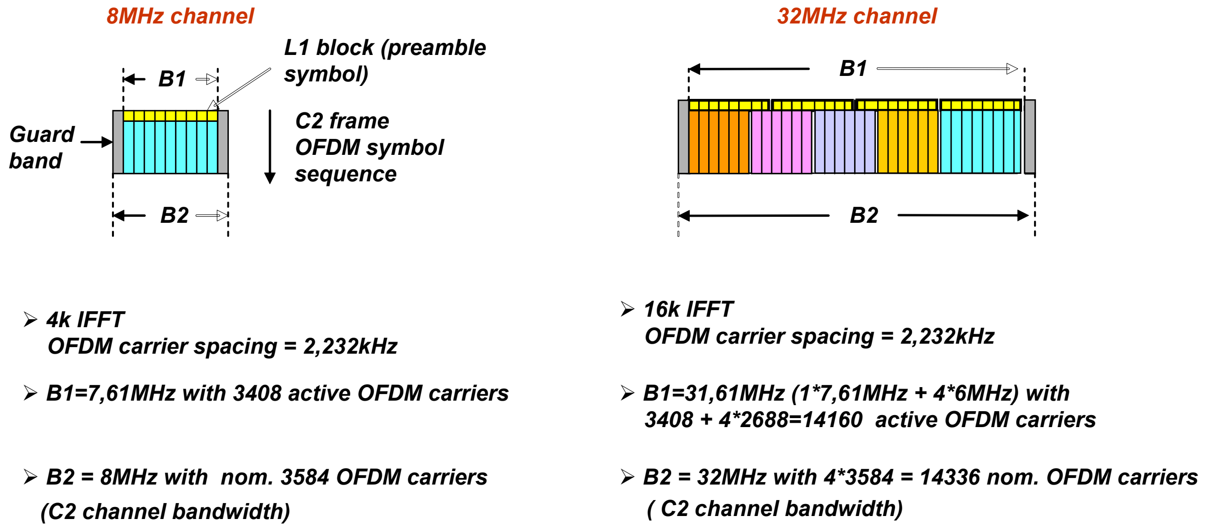


Figure 93: OFDM parameters for an 8 MHz channel and a 32 MHz channel. Data Slice configuration in the 32 MHz channels uses the maximum capacity of 31.61 MHz

Figure 93 shows two practical implementation examples of OFDM parameters for an 8 MHz channel and a 32 MHz channel. In the example of the 32 MHz channel, the maximum capacity of 31.61 MHz (corresponding to 14,160 OFDM carriers) is used.

13.1.3 Practical application example for retransmission from satellite

The block diagram in Figure 94 and Table 11 (table with more detailed transmission parameters) show a practical implementation example for retransmission of 6 DVB-S2 transponders with existing transmission parameters within a 32 MHz wide DVB-C2 channel. This example shows that the optimum transmission capacity can always be achieved, because the DVB C2 transmission rate can flexibly adapted to the input data rate.

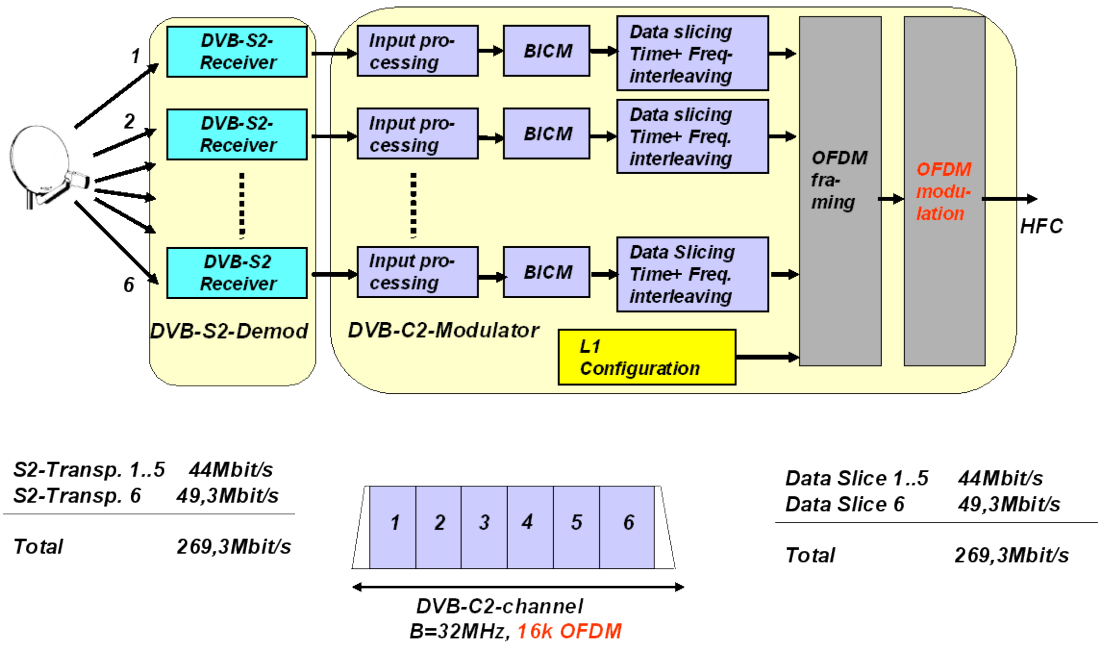


Figure 94: Example for satellite retransmission

Satellite transponders			**DVB-C2 signals**		
Transponders	6		Data Slices	6	
Transponder bit rate in [Mbps]	5 · 44	1· 49.3	Bit rate per Data Slice in [Mbps]	5 · 44	1 · 49.3
			Total DVB-C2 bit rate in [Mbps]	5 · 44 + 1 · 49.3 = 269.3	
Transponder bandwidth	5· 27	33	Channel bandwidth in [MHz]	32	
Modulation	8PSK	QPSK	Modulation (CCM)	1024-QAM	1024-QAM
Code rate	2/3	9/10	Code rate	9/10	9/10
			Number of OFDM carriers (GI = 1/128)*	5 · 2,280	1 · 2,544
Symbol rate	22	27.5	Bandwidth of Data Slice in [MHz]	5 · 5.06	1 · 5.68

* 5 · 2,280 + 1 · 2,544 = 13,944 carriers corresponding to the utilisation of a 16K OFDM; multiple of 24 carriers in each Data slice

Table 11: Example for satellite retransmission, table with more detailed transmission parameters

In the red coloured text of Table 11, the number of necessary OFDM carriers per Data Slice and the overall OFDM carrier number are shown. The number of OFDM carriers per Data Slice is calculated by the formula below (example for the DVB-S2 transponder with 44 Mbps):

$$\begin{aligned} Nc &= (44 \text{ Mbps}) \\ &: [(1/448\ \mu s) \cdot 10\ (128/129) \cdot (9/10) \cdot (58{,}128/58{,}320) \cdot (95/96) \cdot 0{,}99 \cdot (448/449)] \\ &= 2{,}268 \end{aligned} \qquad \text{Eq. 13.1}$$

This formula includes the following factors: carrier spacing (1/448us), order of QAM constellation (10 bps/Hz for 1024-QAM), code rate (LDPC rate = 9/10, BCH rate = 58128/58320), Guard Interval overhead (128/129), pilot overhead (95/96 for scattered pilots and 0.99 for continual pilots), L1 preamble overhead (448/449).

As the number of carriers must be a multiple of 24 (in case Guard Interval GI=1/128 is used), the number of necessary carriers increases to 2,280.

13.2 RF characteristics

In the following chapters, the most important RF characteristics as CINR (taking account of AWGN and intermodulation noise) requirements, spurious emissions, phase noise, Modulation Error Ratio (MER) and Peak to Average Power Ratio (PAPR) will be further described.

13.2.1 CNIR limit, back-off against system level (analogue TV)

As already mentioned above, the new LDPC FEC of DVB-C2 reaches a 6 dB gain concerning QEF ("Quasi Error Free") limit for the minimum required CINR. Therefore DVB-C2 signals using 1024-QAM can be transmitted with the same back-off compared to PAL system level thus with the same level as today's DVB-C signals if 256-QAM is used (see also Figure 95). Consequently, the back-off for a 4096-QAM signal compared to PAL system level can be extrapolated to approximately 0 dB. If in the future PAL channels will be successively switched off, they can be replaced by 4096-QAM DVB-C2 signals. Such a measure will lead to a capacity gain of even 62 % in those channels.

Furthermore, if different CNR values occur at various user outlets, which could happen particularly in larger cable networks, DVB-C2 can further optimise the capacity by applying the feature of variable coding and modulation (VCM) to each individual use outlet. This feature, however, requires the implementation of a return path.

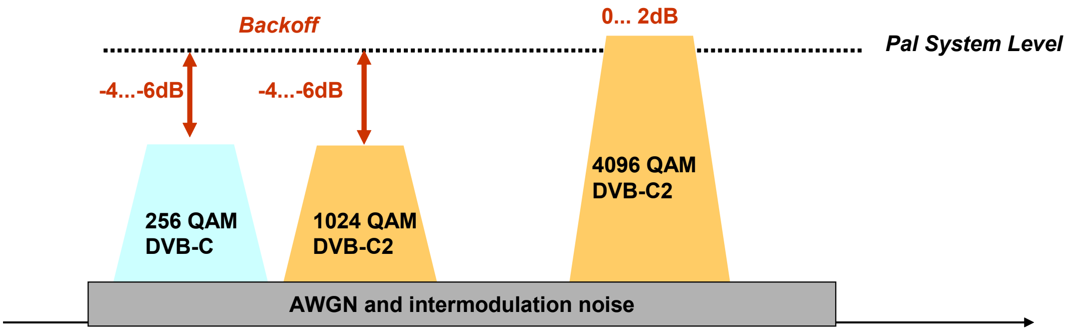


Figure 95: Back-off values in respect of PAL system level

13.2.2 Shoulder attenuation

Figure 96 shows the output spectrum of a DVB-C2 signal. Due to the shown OFDM signal shape the question for required spurious attenuation in adjacent channels arises. Following existing experiences and definitions, all emissions occurring 500 kHz below the lower channel limit and 500 kHz above the upper channel limit must be taken into account.

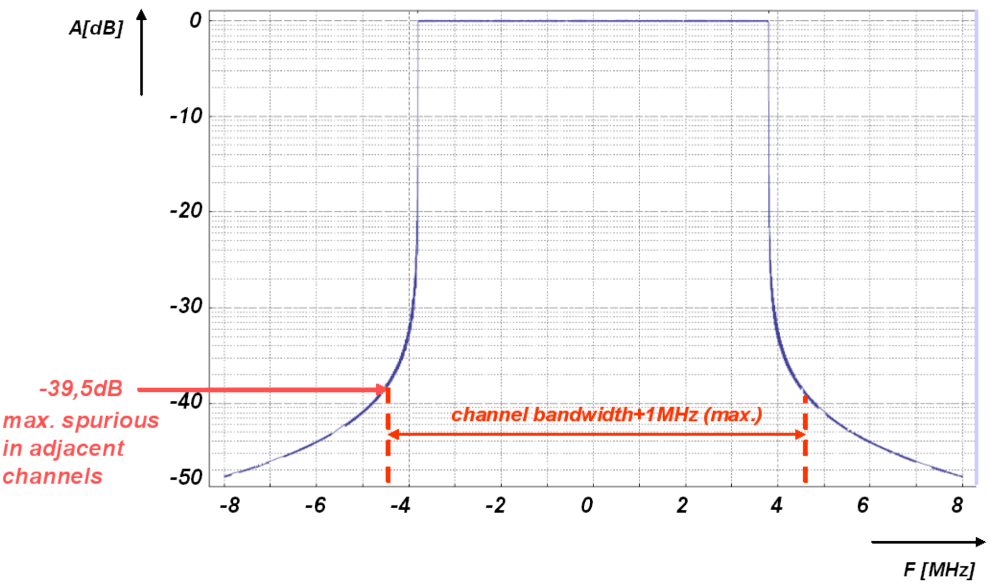


Figure 96: Output spectrum of the OFDM signal of DVB-C2 (4K IFFT, GI=1/128)

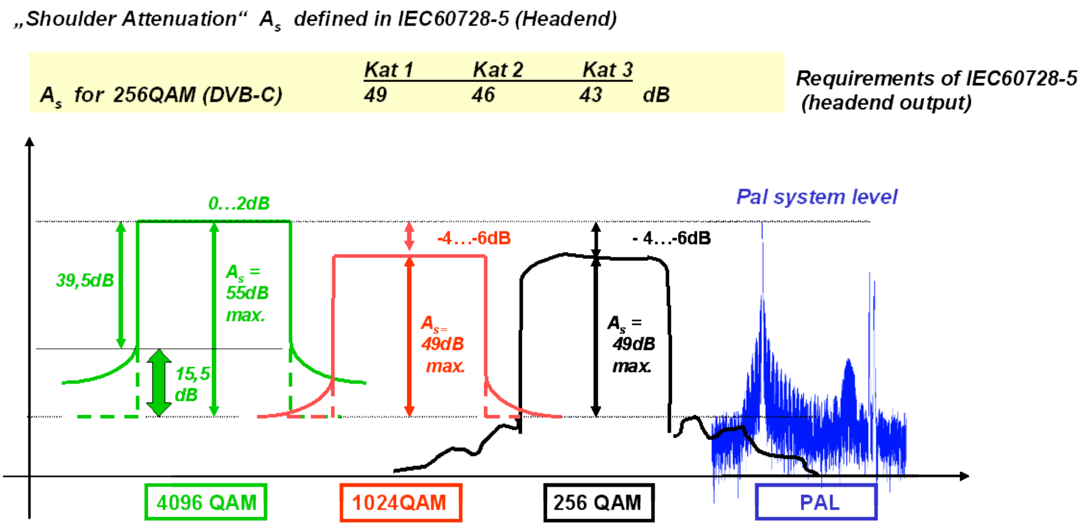


Figure 97: First considerations concerning shoulder attenuation requirements for DVB-C2

For those kinds of emissions into adjacent channels, the shoulder attenuation parameter is defined for example in IEC 60728-5 [13], where the parameters at the headend output are specified. As shown in Figure 97 for 256-QAM (DVB-C) a maximum value of 49 dB is specified for the highest headend category ("Kat 1").

Due to the OFDM signal characteristics and 0 dB back-off of 4096-QAM, the residual signal shoulders of the DVB-C2 signal must be additionally attenuated as indicated by the dashed lines in Figure 97. Assuming a minimum spurious attenuation of 39.5 dB, a filter at the modulator output with an attenuation of min. 15.5 dB must be foreseen to achieve the required shoulder attenuation of 55 dB for 4096-QAM. Of course these first estimations have to be confirmed by more detailed evaluations and measurements, because the signal shoulder characteristic is different between an OFDM and a single carrier QAM signal.

13.2.3 Phase Noise

Phase Noise is a very important parameter for OFDM with increased requirements compared to DVB-C using a single carrier QAM. Figure 98 shows phase noise requirements for OFDM (DVB-T, 64-QAM) derived from a phase noise mask given in terms of a recommendation in EN 301701 (blue coloured line). Consequently the requirements increase by 6 dB (1024-QAM) and 12 dB (4096-QAM) for DVB-C2 has also shown in Figure 98. For comparison, in the frequency band up to 1 kHz, the requirement for DVB-C is only 44 dB (see IEC 60728-5 [13]).

In Figure 98 also the impacts of phase noise on the OFDM signal like CPE (Common Phase Error) and ICI (Inter-carrier Interference) are shown. In frequency regions lower than the OFDM carrier spacing CPE ("Common Phase Error") dominates. CPE assigns the multiplication of RF carrier phase noise with each OFDM carrier. Because this effect overlays at each OFDM carrier in a "common" manner, CPE can be identified using the Continual Pilots and eliminated or strongly reduced by subtracting the phase errors calculated between consecutive pilots. In frequency bands above the OFDM carrier spacing, ICI dominates. ICI behaves like additional random noise and cannot be eliminated.

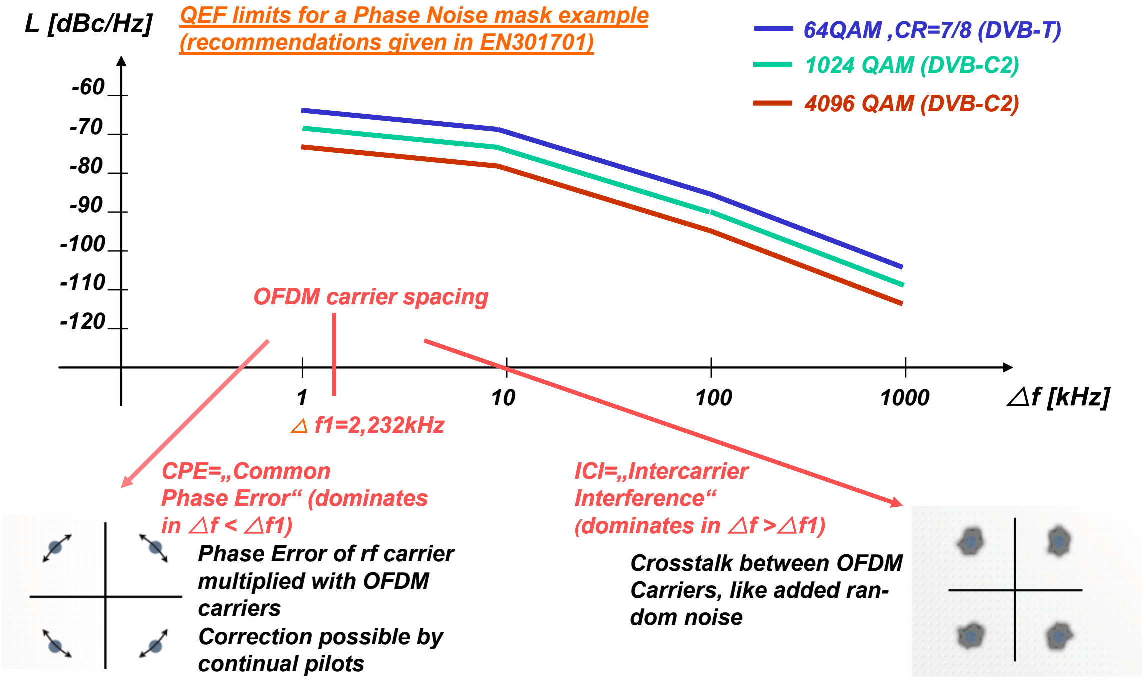


Figure 98: Phase Noise requirements

The conclusion is that for an OFDM signal, phase noise has only relevant effects on ICI, which has also been evaluated in a wide range of OFDM orders with very different phase noise spectra. Therefore tuner manufacturers can calculate the ICI of their tuners by multiplying of phase noise with a weighting function and subsequent product integration as described in Figure 99. Because ICI behaves like AWGN, manufacturers then can estimate the implementation loss of their tuners. This method has been already suggested during DVB-T2 standardisation

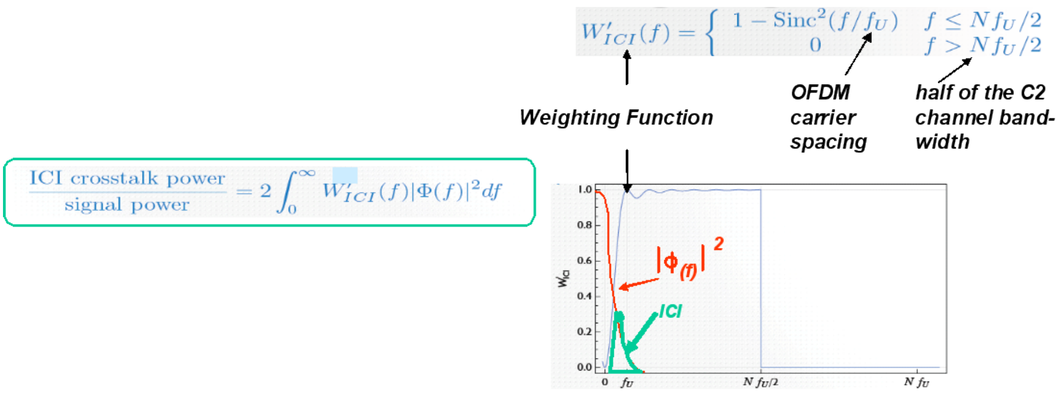


Figure 99: Determination of the impact of phase noise on Inter-carrier Interference (ICI)

13.2.4 Modulation Error Ratio (MER)

The MER values summarize all individual transmission impairments such as AWGN, Intermodulation noise and phase noise as described above. Figure 100 shows the requirements on signal quality defined for the user outlet according to IEC 60728-1 [4] (output parameter specification for the user outlet). Consequently the requirement must be increased by at least 6 dB for 4096-QAM.

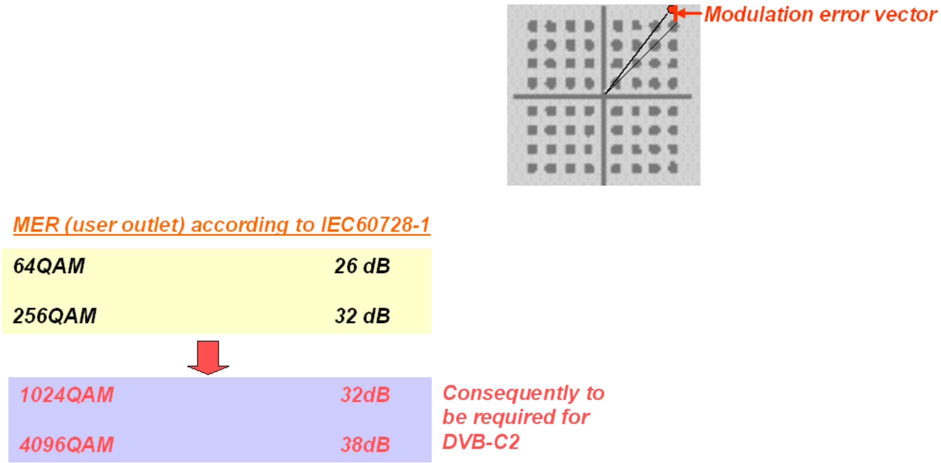


Figure 100: Modulation Error Rate MER

Figure 101 shows the MER values at the user outlet which can be achieved within a cable network by using components and technologies available today.

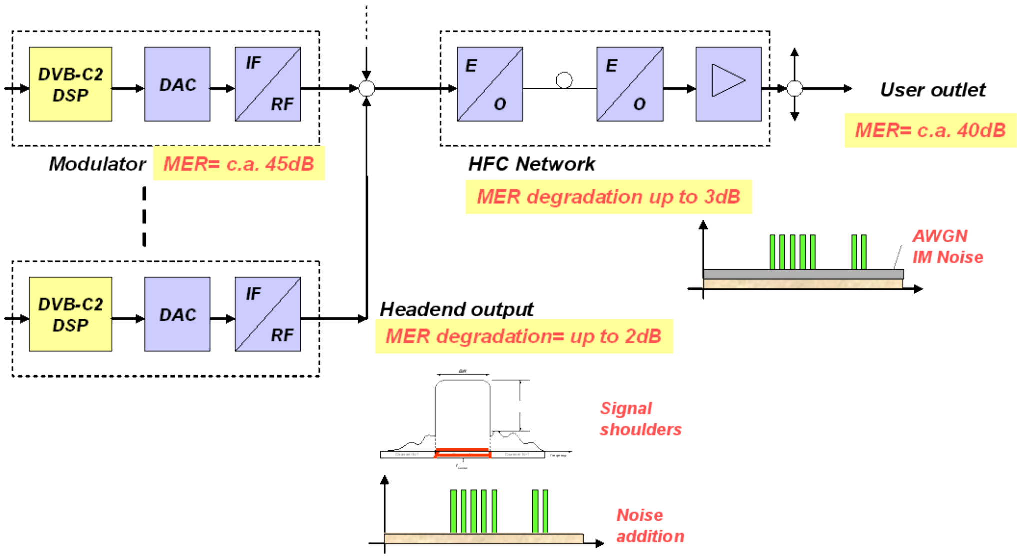


Figure 101: Possible MER values at the user outlet using components and technology available today

13.2.5 Peak to average power ratio PAPR

The transition from single carrier QAM to OFDM requires some examination of the PAPR. 256-QAM (DVB-C) reaches a PAPR value of 7.2 dB, whereas for OFDM a PAPR value of 12 dB must be taken into account. This increase of PAPR has impacts on channel selective components of the cable network installed in headends and on set-top boxes. In the headend, PAPR has effects mainly on required dynamic range of the digital/analogue converter (DAC). But at least state of the art 16 bit DACs easily can reach the required signal shape. On the set-top box side, PAPR principally has effects on the dynamic range of the analogue-to-digital converter (ADC). But at least a state of the art 12 bit ADC can cover the dynamic range requirements, because the relevant input CNR range has a maximum value of about 42 dB including some margin.

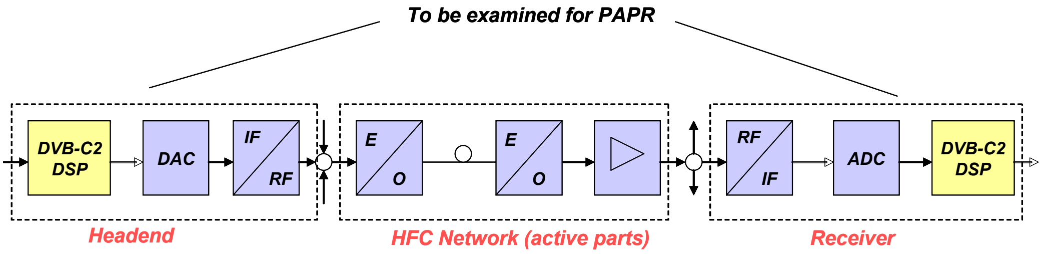


Figure 102: Relevance of PAPR for the transmission chain

For the active parts of the HFC network (optical components, amplifiers) PAPR is considered not have any impact, because the high number of channels transmitted in a cable network are not correlated to each other and, due to the central limiting theorem, the amplitude probability density function (PDF) of the signals converges against a Gaussian distribution independently of the modulation format applied in the cable channels. Nevertheless confirmation for example by measurements of active HFC components with digital OFDM channel load will be necessary.

14 Verification and Validation

Subsequent to the process of preparing the DVB-C2 specification mainly two tasks were worked on by the DVB TM-C2 experts group supported by ReDeSign which were first, specifying the Implementation Guidelines which will support the specification and secondly the Validation & Verification process.

The objective of the Validation & Verification process is summarized within [*Generating DVB-C2 Reference Streams*] as follows:

> *"The main goal of the Verification and Validation (V&V) group is to ensure that the specification is unambiguous and that all companies involved have the same understanding of it."*

This includes the proof of the correct and unambiguous understanding of the DVB-C2 specification by the participating companies with respect to characteristics and implementation of the DVB-C2 components (transmitter side) as well as the review of the DVB-C2 specification and the support of the development of the Implementation Guidelines. Here, the DVB-C2 Validation & Verification process is mainly based on the equivalent DVB-T2 process and its experience.

The basic workflow of the Validation & Verification group is shown in Figure 103. The DVB-C2 system schematic as used in the course of the Validation & Verification process is in general based on the outline of the specification. As a further step, test points have been assigned to the system schematic. The basic work of the Validation & Verification group consists of the definition and refinement of test cases and the subsequent comparison and evaluation of the simulation results related to an individual test case. The latter two steps are repeated for each individual test case and in case the simulation results differ.

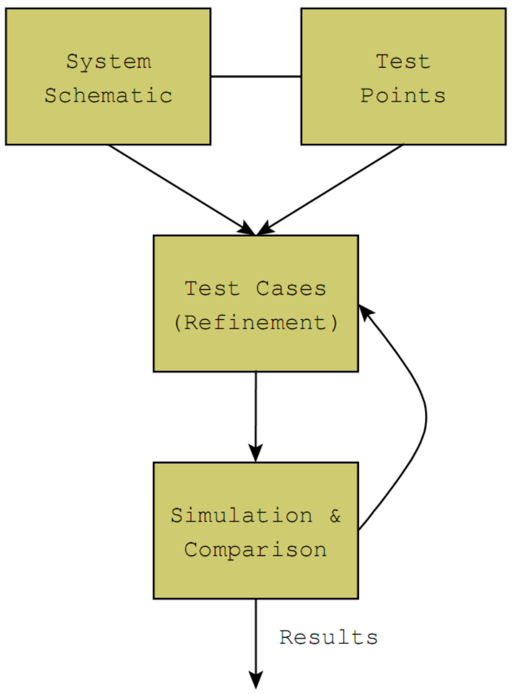


Figure 103: Basic workflow of the Validation & Verification group

- ***<u>Test Points</u>***
  For each distinct component of the DVB-C2 system, a test point is defined and enumerated as depicted in Figure 105. A Test Point can thus be interpreted as a probe sensing the data stream transferred between two connected components resulting in a Test Stream comprising the actual data. This data is subject to a comparison of the Test Streams generated by the individual participants of the

Validation & Verification group. Components which are depending on none or only single system parameters (as selected in the Test Cases) are normally tested only once and left out in subsequent iterations.

- ***Test Cases***
  Test Cases define the parameters of the DVB-C2 system, i.e. a particular configuration, which shall be tested. The main intention is to test preferably all possible scenarios. This includes for instance single and multiple PLP applications as well as single and multiple Data Slices per system. Further configurations will include the optional PLP bundling as well as broad signals exceeding the traditional 8 MHz operation. Concrete test cases are defined in the Validation & Verification internal parameters document [*Generating DVB-C2 Reference Streams (Parameters)*].

- ***Test Streams***
  For each test case and at every test point, a corresponding test stream is generated. This stream comprises the actual data which is the input of the subsequent components of the system. The representation of the stream, i.e. the data format used, is depending on the actual component and the position within the system, respectively. Most common formats used are byte, bit or floating point representations. The stream will be written in a human readable format, to assist a simple checking by the operator, although a binary representation might be more efficient.

For the evaluation of individual components and the complete system, respectively, two different verification strategies are defined: a top-down approach and a bottom-up approach. The top-down approach aims at verifying full modulator configurations. The bottom-up strategy aims at a more exhaustive functional verification of the individual building blocks of the modulator.

- ***Top Down***
  The top-down strategy consists in generating reference streams at each test point, for each modulator configuration. Defining default values is often a source of bugs and should better be avoided. The stream files for all test points can be generated in a single simulation.

- ***Bottom Up***
  The bottom-up strategy consists in verifying each block individually. The verification will start from the input of the modulator and will gradually advance towards the output. Due to the large number of parameters and combinations thereof, verifying all possible combinations is virtually impossible. Instead, for each block, we will only define a very limited number of configurations, also called test cases. The goal is to verify enough of the functionality of each block with as few test cases as possible. The test cases are combinations of parameters that affect the respective block and are defined in a table for each block. Unlike the top-down strategy, where a configuration must define all parameters, a test case only needs to specify those parameters that affect the block under verification, all other parameters taking default values. Test cases can therefore be regarded as partial configurations.

Since the specification describes the DVB-C2 transmission signal of the DVB-C2 system, the Validation & Verification process adverts (only) to the transmitting as well. Thus, the complete system has been broken down into distinct components which basically equal those components as defined and depicted in the specification itself. The complete system schematic as it is used in the course of the Validation & Verification process is shown in Figure 104.

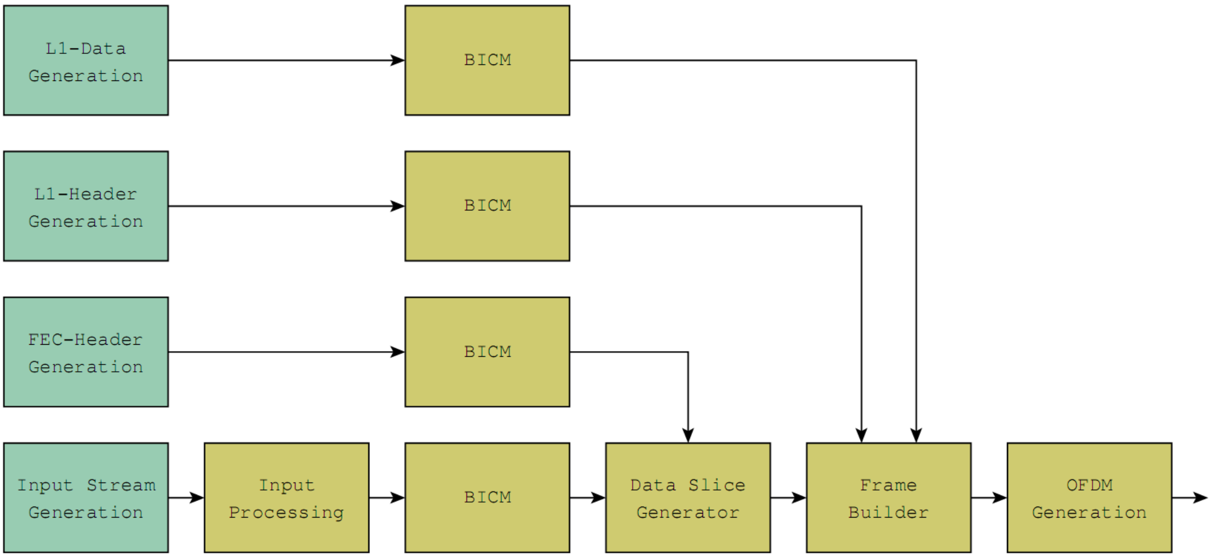


Figure 104: Schematic of the DVB-C2 system at the transmitting side as used in the course of the Validation & Verification process (simplified)

The information being transmitted for which it is subject to different transmission parameters within a DVB-C2 system, i.e. applying diverse error correction or modulation algorithms, can be split into four different paths as shown in Figure 104. These are the paths for the Layer 1 signalling header and data as well as the actual input stream and its optional FEC header insertion. Figure 105 shows exemplarily the Input Processing module of the input stream path with its sub-blocks and the Test Points assigned. The four paths are merged within the Frame Builder to comprise the complete DVB-C2 signal including pilots and reserved tones. The signal is then transformed via the OFDM Generation module into the equivalent analogue baseband signal. The input stream chain as well as the optional FEC header insertion processing is subject to multiplication depending on the actual number of input streams being transmitted by a DVB-C2 system. Thus, the number of data streams being processed by the Frame Builder is variable and limited by the maximum values denoted in the specification. On the contrary, the Layer 1 Signalling information is unique so that the processing chains of the L1 header and L1 data are static and do exist only once per system.

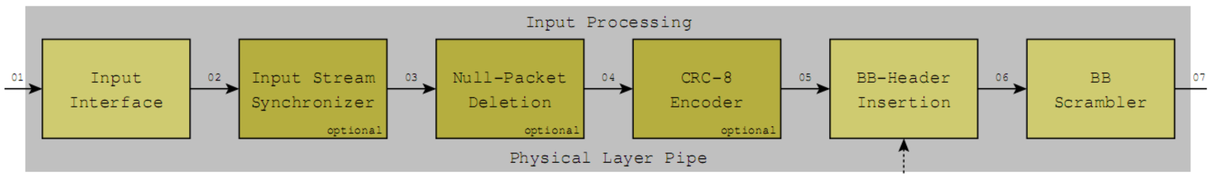


Figure 105: Schematic of Input Processing Module of the input stream path

15 System Aspects and Performance

The preceding chapters describe in great detail the individual techniques used by DVB-C2. At this stage, the reader is also familiar with the DVB V&V process established to validate and verify the techniques and to ensure their unique implementation with regard to standard conformance and vendor interoperability. The remaining chapters focus on the operational and architectural changes which an introduction of DVB-C2 in HFC networks will entail as well as on introduction scenarios and opportunities.

The following explanations of this chapter are results of investigations carried out by the ReDeSign project. They analyze the possibilities for the utilization of DVB-C2 in HFC networks and the potentials provided by the new technology. In the first section, state of the art RF characteristics of HFC networks are explained to an extent important to understand the impacts caused by and the limitations for DVB-C2. Three basic methods are explained which show how DVB-C2 signals could be transmitted in compliance with the existing RF spectrum. The physical characteristics of DVB-C2 can be configured in a very flexible manner, which allows deviating from the traditional cable channel pattern of 6 MHz and 8 MHz, respectively. Instead of a channel number, a new characteristic frequency value is introduced which takes account of the new approach implemented in DVB-C2 called Absolute OFDM. Further to requirements in the frequency domain, the evaluation of future usage scenarios for DVB-C2 requires the consideration of the actual network load in association with the different architectures implemented in the last mile of HFC networks. A more detailed investigation of which network load could be transmitted through which topology is described in the subsequent chapter 16.

15.1 HFC RF spectrum considerations

HFC networks carry different kinds of broadcast and broadband downstream signals from headends to customers as well as broadband upstream signals in reverse direction. Next to analogue FM radio, analogue and digital TV as well as DOCSIS signals are transmitted in an RF spectrum ranging from 5 to maximal 65 MHz (upstream) and from 87.5 MHz to 862 MHz (downstream). Current investigations to extend the two frequency bands to higher frequencies are underway but not described hereafter. A Frequency Division Multiplex (FDM) system is applied to allow a distinct signal separation at the receiving end. Because of

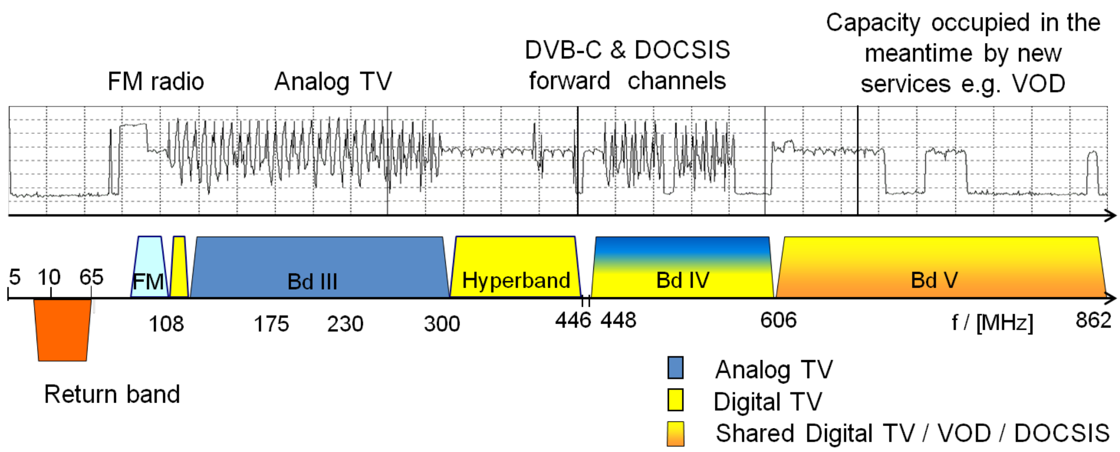


Figure 106: Example of occupation of RF spectrum in an 862 MHz HFC network

these arrangements, RF spectrum diagrams are well suited for an analysis of transmission scenarios in a network. Figure 106 shows an example of such an RF spectrum. The spectra of the individual signals traditionally transmitted e.g. FM radio, analogue PAL, DVB-C and (Euro)DOCSIS are arranged in the FDM manner with an equidistant channel pattern provided by the network. Frequency gaps are noticeable in the higher frequency sub-bands, particularly in Band IV and V (see Figure 106). In many HFC networks these sub-bands have been occupied to a large extent and are used today for the introduction of new services such as HDTV, VOD, and Catch-up TV etc.

The introduction of DVB-C2 signals in such a scenario needs to take account of a number of technical requirements; however, the most important one certainly is the demand to prevent any reduction of technical quality of the traditional signals. This important requirement can be fulfilled if DVB-C2 signals comply with the transmission parameters defined by the relevant standards published by IEC and CENELEC [4]. In particular channel frequencies and power levels are of interest which means that a DVB-C2 signal is to be injected in the existing channel raster at a defined power level. For the following discussion it is assumed that a sub-band of the RF frequency spectrum is occupied with signals as indicated in Figure 107 a) by means of an example. A simple means introducing DVB-C2 is a one-to-one exchange of an individual signal such as analogue TV or DVB-C with a DVB-C2 signal while keeping the signal power at the dedicated level. The resulting RF spectrum of such a simple example is depicted in Figure 107 b). Two DVB-C signals transmitted in adjacent channels and one analogue TV signal are replaced each by a DVB-C2 signal. Figure 107 c) shows the possibility to transmit a wide-band DVB-C2 signal of a bandwidth equivalent to two cable channels (e.g. 16 MHz in Europe, 12 MHz in the U.S.). The benefit arising from the injection of wide-band signals in two or more adjacent channels consists of an increase of spectral efficiency caused by an active use for data transmission of the frequency Guard

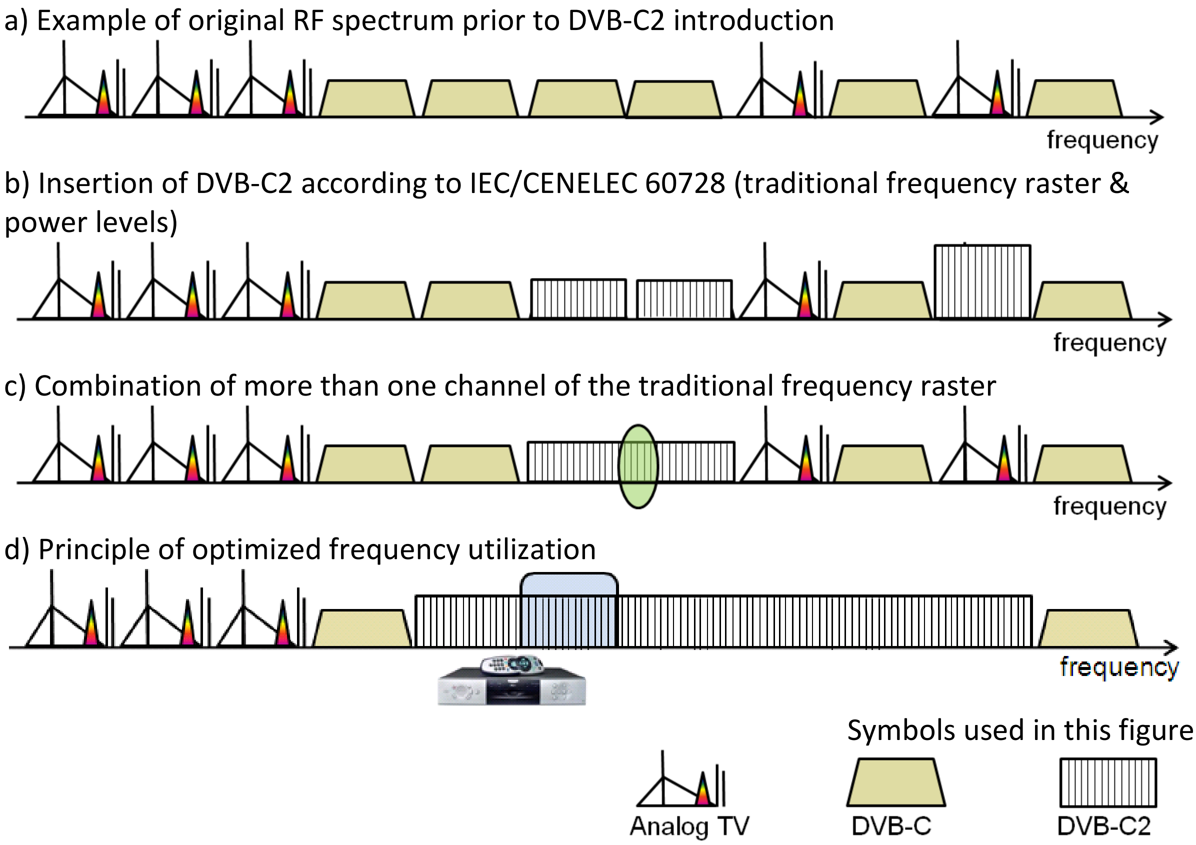


Figure 107: Opportunities for RF spectrum utilization

Band traditionally applied between two adjacent signals. In Figure 107 c) such frequency band is marked by the coloured ellipse. Compared to this example of a combination of two channels, the spectral efficiency can be further accelerated by either transmitting signals of even wider bandwidths and combining a higher number of channels or by injecting individual DVB-C2 signals adjacent to each other and without making use of any frequency Guard Band. Such a close packet signals line-up can be achieved if the signals comply with the pre-requisites defined for the special case of an Optimized Frequency Utilization. The pre-requisites are illustrated further down in this chapter. An example of the resulting RF spectrum is depicted in Figure 107 d).

As described already in various chapters above, DVB-C2 receivers using tuners with a traditional receiving bandwidth (of 8 and 6 MHZ, respectively) are capable of receiving individual Data Slices or bundles of Data Slices transmitted in related frequency bands. Guided through the DVB-C2 Signalling Information, the CPE will be tuned to the Data Slice which carries the service of interest, e.g. an HD program.

15.2 Optimized Frequency Utilization

The pre-requisite for the application of an Optimized Frequency Utilization is the establishment of a mode requiring the full synchronization of all DVB-C2 signals being part of a DVB-C2 ensemble. The synchronization has to be created in both time and frequency domain. It is necessary to ensure that the physical behaviour of all synchronized DVB-C2 signals of the entire ensemble is identical with the behaviour of a single wide-band DVB-C2 signal. In fact the orthogonality of the OFDM sub-carriers applied in the frequency domain has to be guaranteed not only within an isolated OFDM symbol but also among the sub-carriers being part of each adjacent OFDM symbol and thus of each OFDM symbol contributing to the ensemble of DVB-C2 signals. This synchronization requirement needs to be fulfilled at least during the active symbol duration (excluding the temporal Guard Interval). Therefore the DVB-C2 signals of the ensemble have to be synchronized in time domain as well.

The spectrum of such an ensemble is shown in Figure 107 d). While the bandwidth of a single DVB-C2 signal can be assigned in a very flexible manner, the bandwidth of such an ensemble of DVB-C2 signals can vary significantly and has only a lower limit which is equal to the minimal bandwidth of a single DVB-C2 signal.

15.3 Spectral implications of the Absolute OFDM mechanism

The introduction of the so-called Absolute OFDM concept (see above and sub-chapter 10.4) entails the definition of a new set of RF parameters corresponding to the traditional RF channel numbering: The OFDM Sub-carrier Index and the L1 Channel Number. As shown in

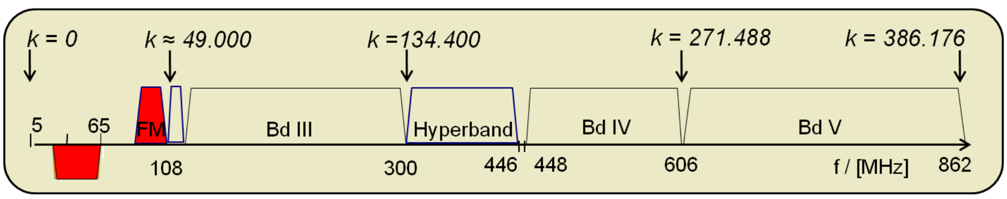


Figure 108: Definition of Absolute OFDM sub-carrier grid: number of absolute carrier k and respective frequency in MHz

Figure 108, the entire RF frequency band used for up- and downstream transmission in HFC networks is subdivided for this purpose in sub-bands of the OFDM sub-carrier bandwidth f_{SC} = 1/448 MHz ≈ 2,232 Hz and parameterized with a running index k. The Sub-carrier Index k starts with a value of 0 indicating the frequency sub-band at DC (0 Hz) and ends at a value of 386,176 which corresponds with a frequency of 862 MHz. The interrelationship between the index k and the medium frequency of the related frequency sub-band is given by the formula:

$$f_{sc,k} = k/_{448} MHz \quad with \quad f_{SC,\min} = 0MHz \quad and \quad f_{SC,\max} = 862MHz \qquad \text{Eq. 15.1}$$

As explained already in previous chapters (e.g. sub-chapter 10.3), L1 Blocks are aligned to a fixed 7.61 MHz frequency grid which corresponds to 3408 OFDM sub-carriers. The theoretical concept arranges for the first L1 Block starting at 0 Hz whereas the subsequent L1 Blocks starting at sub-carrier indexes of a multiple number of 3408. They can be calculated as follows:

$$k_{L1-Start} = n \cdot 3{,}408 \quad for \quad n = 0,1,2,...,113 \qquad \text{Eq. 15.2}$$

Assuming that the FM radio band will be continuously used for analogue audio transmission, the first entire L1 Block transmitted starts at $k_{L1\text{-Start, min}}$ = 15 corresponding to 114.1 MHz. However, parts of the L1 Block could also be transmitted below this frequency but above 108 MHz which is the upper frequency limit of FM radio. A protection distance between DVB-C2 and the FM signal needs to be taken into account when introducing DVB-C2 in these frequency bands. The required protection ratio is currently investigated by ReDeSign and under standardization by CENELEC and IEC.

15.4 Examples of transmission impairments

This subchapter explains some results of transmission impairments received by computer simulations carried out by Institut für Nachrichtentechnik of Technische Universität Braunschweig in the course of the ReDeSign activities. First the simulation platform is introduced; Individual transmission disturbances selected by means of examples are analyzed afterwards (for further information see also [7]).

15.4.1 Simulation platform

Performance investigations of DVB-C2 transmission through cable networks were carried out by means of computer simulations based on a comprehensive simulation platform. The platform consists of a transmitter block covering the signal processing complying with DVB-C2 and a respective receiver block executing the inverse processing. A sophisticated cable channel model [3] was developed to investigate the transmission performance of the new technology. The model considers transmission disturbances typically occurring in cable networks. In this sub-chapter, simulation results of AWGN, narrow-band interference and impulse noise are used to explain the functioning and performance of DVB-C2 as well as the impact of individual processing units installed at the receiving end. A high-level diagram of the simulation platform is shown in Figure 109.

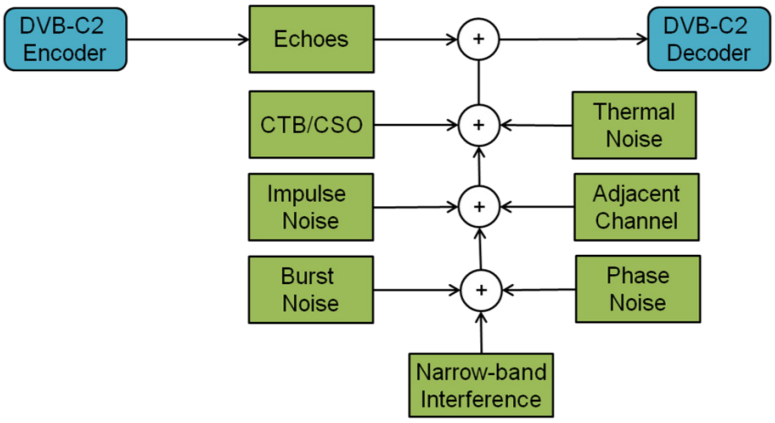


Figure 109: Block diagram of DVB-C2 simulation platform

In the following, the impact to a DVB-C2 transmission caused by narrow-band interference and impulse noise is explained by means of examples. Each of the two kinds of interference has a typical character which is helpful to understand the effect of the time and frequency interleaving, respectively, implemented in DVB-C2

15.4.2 Performance simulation in AWGN channel

A fundamental disturbance occurring in all real transmission systems is thermal noise which is generated by various network elements such as amplifiers appearing along the entire cable network including transmitting and receiving ends. Thermal noise is generally considered in a simulation by means of additive white Gaussian noise which is also the case in the simulation described by this article. A basic AWGN channel implementation allows, for instance, a comparison of different transmission systems in terms of spectral efficiency as well as with the theoretical limit defined by the Shannon Theorem [6]. The diagram of Figure 110 shows the simulation results allowing such a comparison with the DVB-C system. DVB-C is the predecessor to DVB-C2 and was developed in 1994. Furthermore, the diagram shows that the different DVB-C2 LDPC code rates in combination with the modulation constellations allow an optimization of the spectral efficiency with a granularity of some 2 dB throughout the SNR range between 10 and 35 dB. The gain in terms of robustness compared to DVB-C is almost 7 dB. This SNR benefit, for example, is achieved when switching from a DVB-C 256-QAM to DVB-C2 using the same constellation in combination with a code rate of 5/6. The horizontal arrow taking course from right to left between the two respective points in the diagram is indicating this benefit. The minimal advancement in terms of spectral efficiency is given by switching from DVB-C 256-QAM to DVB-C2 1024-QAM using a code rate of 9/10. In this case, the spectral efficiency increases from 6.4 bps/Hz to 8.7 bps/Hz which is equivalent to some 35 % (see vertical arrow between the two respective points in the diagram). In contrast to DVB-C, DVB-C2 can further increase the spectral efficiency up to more than 10 bps/Hz if cable channels support SNR values higher than the 30 dB required to transmit a quasi error free DVB-C signal using 256-QAM. The gain received from switching from DVB-C 256-QAM to DVB-C2 4096-QAM is close to 60 %. At this stage it should be noted that the following DVB-C2 parameters impacting the results were selected for the simulation: channel bandwidth of 32 MHz, OFDM Guard Interval of 1/128, and OFDM pilot density of 1/96.

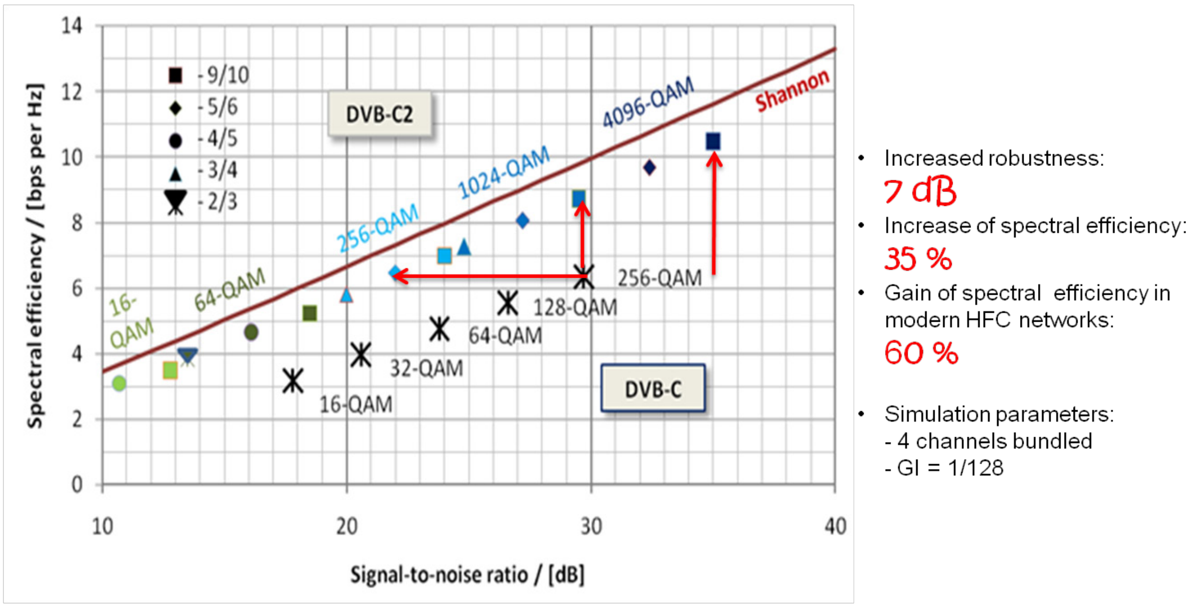


Figure 110: Spectral efficiency as a function of signal-to-noise ratio; comparison between DVB-C2 and DVB-C using the Shannon Limit as reference

15.4.3 Narrow-band interference

Narrow-band interference was simulated with a bandwidth of 100 kHz and with a power level which was chosen to be high compared to the portion of the DVB-C2 affected. This scenario defines some kind of worst case for simulations of narrow-band interference since firstly the chosen bandwidth of 100 kHz is rather wide for this interference type and secondly the DVB-C2 subcarriers impaired by the interference were completely destroyed. In practice, such extremely high power levels of interferer can cause additional non-linear effects resulting, for instance, in an override of the RF or ZF stages in receivers, which could entail further disturbances affecting the entire DVB-C2 signal. However, these non-linear effects are dependent on the implementation of the receiver unit itself and, thus, were not considered in the simulation. In fact, in the simulation it became visible that the narrow-band interference generates a strong error pattern focused at the affected DVB-C2 sub-carriers. The Frequency De-interleaver at the receiving end of the simulation platform distributed the error pattern equally and quasi randomly over an entire Data Slice and, in the case simulated, over a bandwidth of 7.6 MHz. This effect of dispersing the energy of the erroneous bits through an entire Data Slice allowed the subsequent forward error correction (LDPC and BCH) to correct the destroyed bits completely. In case such strong narrow-band interference occurs, DVB-C2 allows to simply ignore a dedicated number of adjacent sub-carriers, called notching. This kind of notching can be applied in this scenario.

Interesting is the knowledge of the performance degradation which is caused by a narrow-band interference in case it is added to an additive noise component. Simulation results are shown in Figure 111. The degradation between the curve indicating the performance of DVB-C2 in an AWGN channel (solid line) and the two curves involving additional narrow-band interference (dotted lines) is equal to a fraction of a dB. Note that the total abscissa shown in the diagram reflects an SNR difference of only 0.6 dB. For sake of completeness it should be mentioned that a DVB-C2 256-QAM was used for the simulation.

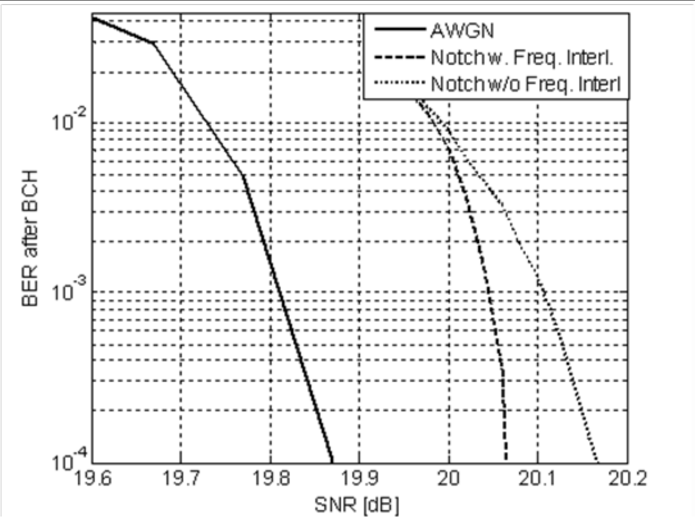


Figure 111: Remaining BER of DVB-C2 (256-QAM) after FEC (LDPC & BCH) generated by AWGN overlaid by narrow-band interference; frequency interleaving switched on and off

15.4.4 Impulse noise interference

A complete opposed disturbance characteristic to narrow-band interference is caused by impulsive noise occurring in terms of temporal impulses. It is assumed that each impulse event has a rather short life cycle and a wide bandwidth. As typical source for impulse noise, the European GSM system was selected by means of an example. The duration of an eliminated GSM impulse event is some 550 µs which is in the order of a single OFDM symbol and, thus, is short compared to a DVB-C2 Frame composing a number of OFDM symbols. Each impulse event is followed by a large off-time period during which no bursts are sent. The bandwidth of the bursts is larger than the one of a DVB-C2 Data Slice. It is also assumed that the power level of each pulse is much higher than the power level of an OFDM symbol which causes the complete destruction of an entire OFDM symbol during the active burst event. That is why the possibility to correct the transmitted bits does not depend on the actual QAM constellation. In contrast it largely depends on the chosen LDPC code rate. The fact that the positions of the destroyed OFDM symbol are known to the receiver and, thus, can be marked accordingly allows the LDPC decoder to apply the puncturing mechanism. In case a good code rate is used with the number of parity bits being higher than the number of punctured bits affected by an impulse, the receiver is able to correct the erroneous bits entirely. The error correction performance of DVB-C2 for this scenario is given by Table 12. The performance increases with increasing the depth of the Time De-interleaver. Applying a depth of 16 OFDM symbols, even the weakest code rate of 9/10 can be used to correct all bits disturbed.

15.5 HFC capacity estimations

With the knowledge gained by the simulations and investigations regarding Optimized Frequency Utilization, explained in the above sub-chapters, calculations were carried out to estimate capacity gains which an introduction of DVB-C2 may generate.

For a first estimation of capacity available in cable networks for digital signals, data were taken from the results of a survey which the ReDeSign project conducted among European cable operators in 2008. From the responses of the cable operators representing almost a third of the entire European operators' community, in terms of numbers of operators and subscribers served by their networks, average scenarios were derived and dealt with as the basis for the following calculations [5], [8].

Depth of Time Interleaver	FEC code rates required for QEF reception
1 OFDM symbol	no code rate
2 OFDM symbols	no code rate
4 OFDM symbols	2/3
8 OFDM symbols	2/3, 3/4, 4/5, 5/6
16 OFDM symbols	all code rates

Table 12: DVB-C2 code rate required for quasi error-free (QEF) reception in dependence of the interleaving depth

Scenario 1 describes an average service portfolio transmitted through cable networks today. The number of occupied channels was chosen to be 95 – corresponding with a fully loaded frequency spectrum in downstream – with 40 channels being used for the provision of analogue TV services. Further 44 channels were assigned to digital TV signals and correspond to a total capacity of 1.9 Gbps of which 0.9 Gbps were used for STDV using DVB-C with 64-QAM and 1 Gbps for HDTV and VOD services based on DVB-C with 256-QAM. 11 DOCSIS channels provide High Speed Internet (HSI) and other IP based services. Respective bit-rate figures are given in Table 13.

Service	Technology	No of channels x constellation	Digital capacity
Analogue TV	PAL, SEC.	40	n.a.
SDTV	DVB-C	24 x 64-QAM	0.9 Gbps
HD & VOD	DVB-C	20 x 256-QAM	1.0 Gbps
HIS/IP	DOCSIS	11 x 256-QAM	0.5 Gbps
Total		95 channels	2.4 Gbps

Table 13: Estimation of digital capacity for a today's channel line-up

In Scenario 2, the number of analogue TV signals was reduced to 25. The freed spectrum was used either to introduce new services such as HD and VOD via DVB-C2 or to migrate the existing services to the new technology. Assuming that a 4096-QAM could be applied, the 120 MHz frequency slot (equivalent to fifteen 8 MHz channels) makes an additional digital capacity of approximately 1.25 Gbps available. This capacity allows converting the entire HD and VOD services from 20 DVB-C channels to 15 DVB-C2 channels while providing opportunities for service extensions through the introduction of new services.

In a next step, the 20 DVB-C channels occupied with 256-QAM signals were upgraded to DVB-C2 using 1024-QAM. This step increased the capacity of the corresponding 160 MHz band from 1 Gbps to some 1.4 Gbps and allowed a migration of the SDTV service to DVB-C2. Since the composite bit rate of all SDTV programs was smaller than the capacity available, either the SDTV service or another TV service using DVB-C2 technology could be extended by introducing new programs. Eventually, the spectrum freed from the 24 SDTV channels could be entirely used by DOCSIS – for instance for the introduction of an IPTV service – while enlarging the DOCSIS capacity from 0.5 Gbps to 1.4 Gbps (see Table 14). The QAM constellations of DVB-C2 were chosen according to the diagram in Figure 110 considering an SNR value which was required for a DVB-C transmission.

Service (technology)	Technology	No of channels x constellation	Digital capacity
Analogue TV	PAL, SEC.	25	n.a.
SDTV	DVB-C2	15 x 4096-QAM	1.3 Gbps
IPTV	DOCSIS	24 x 64-QAM	0.9 Gbps
HD & VOD	DVB-C2	20 x 1024-QAM	1.4 Gbps
HIS/IP	DOCSIS	11 x 256-QAM	0.5 Gbps
Total		95 channels	4.1 Gbps

Table 14: Estimation of digital capacity for a system using DVB-C2 for TV and DOCSIS for HIS/IP services

The final scenario used DVB-C2 as an integrated portion of DOCSIS for downstream transmission. If the remaining 25 analogue TV channels were replaced by DVB-C2, the entire downstream channel line-up would be filled with DVB-C2 signals. Assuming that again a 4096-QAM constellation could be used in these former analogue TV channels and a 256-QAM DVB-C2 signal in the former SDTV channels, the spectrum would be occupied as indicated by Table 15. The entire digital capacity of the cable network reached a figure of almost 6.9 Gbps.

Service (technology)	Technology	No of channels x constellation	Digital capacity
Band I	DOCSIS & DVB-C2	40 x 4096-QAM	3.4 Gbps
Band II		24 x 256-QAM	1.3 Gbps
Band III		31 x 1024-QAM	2.2 Gbps
Total		95 channels	6.9 Gbps

Table 15: Estimation of digital capacity provided by an integrated DVB-C2/DOCSIS system utilized in the entire cable downstream spectrum

15.6 DOCSIS integration issues

The rough estimations carried out in the last sub-chapter demonstrate how significant the capacity assigned to digital signals can be increased when introducing DVB-C2 and finally integrating the technology in DOCSIS for downstream transmission. For example, the combination of twelve 8 MHz channels to a single 96 MHz DVB-C2 downstream pipe using 4096-QAM would allow the establishment of a 1 Gbps downstream connection mode via HFC. Such enormous capacity per connection has of course many operational side effects. For instance, it paves the way for the creation of an integrated solution for full service provision based on IP. Bandwidth could be dynamically assigned to different service types in an easy manner. On the other hand, the integration of DVB-C2 in DOCSIS requires some additional specification and standardization work. For instance, CMTS and Edge Resource Manager have to be aware of the advanced capacity figures provided by the new technology and also timing and synchronization between upstream and downstream needs to be established within the existing DOCSIS requirements

16 Network Planning

When deploying DVB-C2, it will be attractive to apply a high signal level because a high signal level permits the use of 4096 QAM modulation mode corresponding to a bit rate of more than 80 Mbps. However, raising the signal level conceivably may degrade the signal quality of the digital and analogue signals due to the non-linear nature of the amplifiers and optical transmitters.

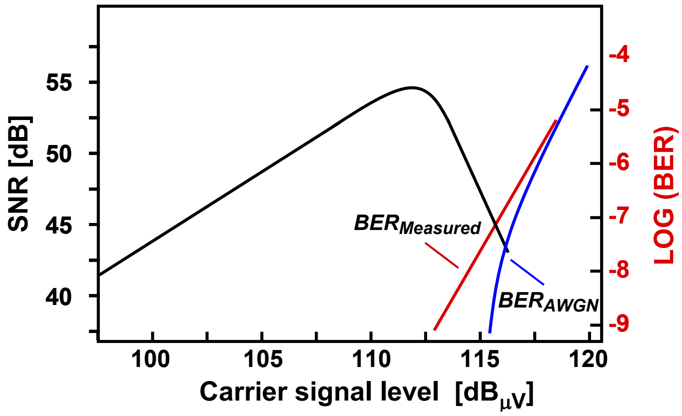


Figure 112: Qualitative illustration of the non-linear behaviour of an amplifier in case of a full digital load.

As a mind-setter, Figure 112 shows the degradation of the digital carrier signal when raising the composite load applied to an amplifier. For a low carrier level, the noise level is dominated by the thermal noise level, and increasing the carrier level results in a proportional increase of the signal-to-noise ratio (SNR) and a quasi error-free transmission (before forward error correction, FEC), as reflected by the slope +1 of the SNR curve and a bit error rate (BER) $<10^{-10}$. When raising the carrier level beyond a specific threshold the non-linear behaviour becomes notable and the slope of the SNR curve is gradually bend into a steep decline. In addition, the generation of bit errors increases above a bit error rate of 10^{-10}.

The BER indicated in Figure 112 refers to the BER before FEC, and as such applying FEC will extend the quasi error free operational range with a few dB to a higher carrier level. However, in case of a network with a mixed analogue digital network load, the quality of the analogue signals must be taken into consideration as well.

In this chapter an analysis of this DVB-C2 RF planning issue is given. On the one hand, a sufficiently high DVB-C2 signal level is needed to warrant good reception. On the other hand, a too high DVB-C2 signal level would raise the distortion signal levels thus causing the degradation of the signals.

16.1 The nature of the non-linear distortion products

In agreement with signal theory, non-linear behaviour is described by a Taylor expansion:

$$y(t) = a_1x(t) + a_2x^2(t) + a_3x^3(t) + a_4x^4(t) + a_5x^5(t) + \ldots \qquad \text{Eq. 16.1}$$

With the aid of the Price Theorem and provided that $x(t)$ is a Gaussian signal, this time-domain response function can be restated in terms of a frequency-domain description:

$$Y(\omega) = A1X(\omega) + A2X(\omega)\otimes X(\omega) + A3X(\omega)\otimes X(\omega)\otimes X(\omega) + A4X(\omega)\otimes X(\omega)\otimes X(\omega)\otimes X(\omega) + \ldots \quad \text{Eq. 16.2}$$

with $Y(\omega)$ and $X(\omega)$ the power spectral density of the signals $y(t)$ and $x(t)$ respectively [9].

Thus, in the frequency domain, the non-linear behaviour results in the generation of intermodulation (IM) products, where the spectrum of the distortion signal is given by the convolution of the composite input spectrum.

Current network loads comprise digital signals with a broadband nature and analogue TV signals. The signal of the latter is dominated by the non-modulated carrier, and as such the analogue TV signals can be approximated as a narrowband signal. Because of this mixed analogue-digital load, the spectrum of the IM signal at the output of the component exists of:

- Narrowband distortions; the well-known CSO/CTB cluster beats (these are associated with the intermodulation between analogue TV signals),
- Broadband distortions

The broadband distortion signal is generated by either the intermodulation of the analogue TV signals and digital signals or by intermodulation of digital signals with digital signals. From the viewpoint of signal theory, the broadband distortion signal has a random noise nature[4].

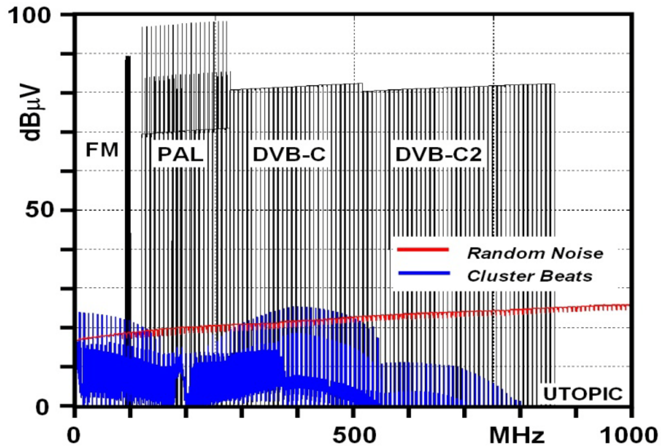


Figure 113: Spectrum of cable signals and intermodulation products for a cascade of amplifiers with a mixed load of PAL and digital carriers as "measured" at an amplifier output port using a spectrum analyzer with 300 kHz bandwidth resolution. The figure is obtained from simulation using a 2nd and 3rd order component model.

Figure 113 shows the spectrum of the narrowband (blue, composite cluster beats) and the broadband distortion signal (red, random noise) for a component with a mixed load. This figure is obtained from a simulation. Because of the relatively low digital carrier level, the intermodulation between digital carriers is negligible; however, intermodulation between TV signals and digital carriers is seen as an 8 MHz ripple reflecting the power spectral density of the composite signal of the digital carriers.

Apart from the random noise and narrowband cluster beats, various studies report the occurrence of impulse and/or burst noise. Considering the component model of Eq.1 or Eq. 2, the occurrence of such impulse events is not straightforwardly evident. To properly understand this issue, we have measured and studied time-domain samples of the RF distortion signal. Using a set-up with a frequency down converter and a fast capture card, we have measured the baseband distortion signal in an 8 MHz RF channel for different cable amplifiers with a composite load of 95 DVB-C carriers. From the real-time samples of the

[4] For very high composite digital loads, impulse events are generated as well, see further on in this section.

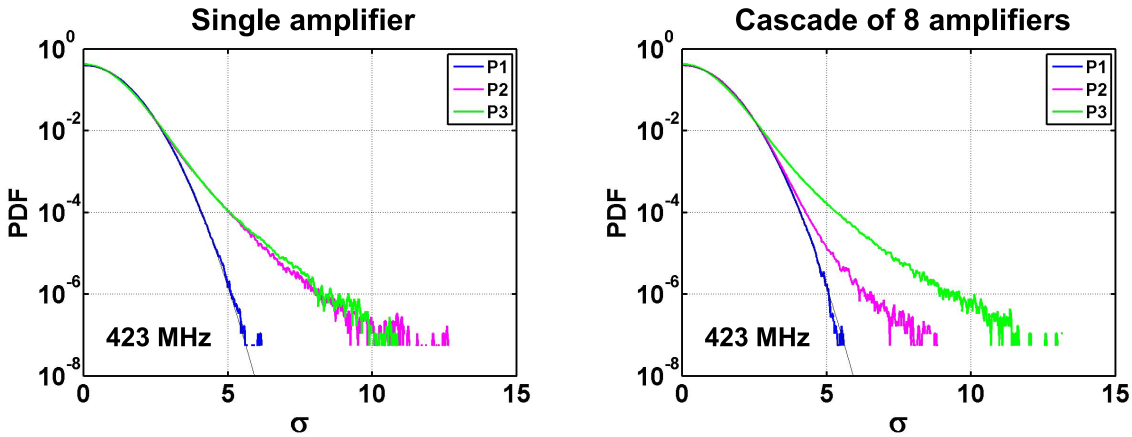


Figure 114: PDF of the distortion signal of a single amplifier (left window) and a cascade of 8 amplifiers (right window) with a composite load of 95 digital carriers. The PDFs are recorded for a carrier level that corresponds with the maximum of the SNR curve (P2), a carrier level 3dB below P2 (P1) and 3 dB above P2 (P3). All curve are normalized to the average distortion signal power (σ). The grey curve shows the PDF for random noise (Gaussian distribution). The deviations at P2 and P3 are associated with impulse events.

baseband distortion signal, we have calculated the probability density function (PDF) which shows the statistical occurrence of the distortion signal level. Some typical results for a single amplifier and for a cascade of 8 amplifiers and for 3 different carrier levels (P1, P2 and P3) are shown in Figure 114. Comparable results were obtained for different amplifiers and for a mixed analogue digital load. In these measurements, the carrier level P2 corresponds with the maximum of the SNR curve whereas P1 and P3 were chosen respectively 3 dB below and 3 dB above P2. Thus, for P1 the non-linear effects should be negligible and thermal noise dominant. At the carrier levels P2 and P3, the non-linear distortion products should contribute notably (P2), if not dominant (P3). All PDFs are normalized to the average distortion signal power, and as a reference, a Gaussian distribution associated with a random process is indicated in the figure.

Figure 114 shows a Gaussian PDF of the distortion signal for a (low) carrier level P1 and an extended tail in addition to a Gaussian distribution for the higher carrier levels P2 and P3. Further analysis of the time domain samples of the distortion signal showed that this tail corresponds with the occurrence of impulse events with a random occurrence in time (no bursts or trains of impulses) and a typical duration of about 100 ns,[5] [3]. In addition, Figure 114 reveals that the impulse events have a mild nature; the signal level is roughly 10 times the average distortion signal level (σ), or about 10 dB above the distortion signal level, or about 3-4 dB above the signal peaks found in a random distortion signal with the same average power. In Figure 112 we have indicated the expected BER in case of additive white Gaussian noise (BER_{AWGN}), which is lower than the measured BER ($BER_{Measured}$). This observation is consistent with the occurrence of such (mild) impulse events.

[5] The 100 ns duration is related to the 8 MHz bandwidth of the captured distortion signal.

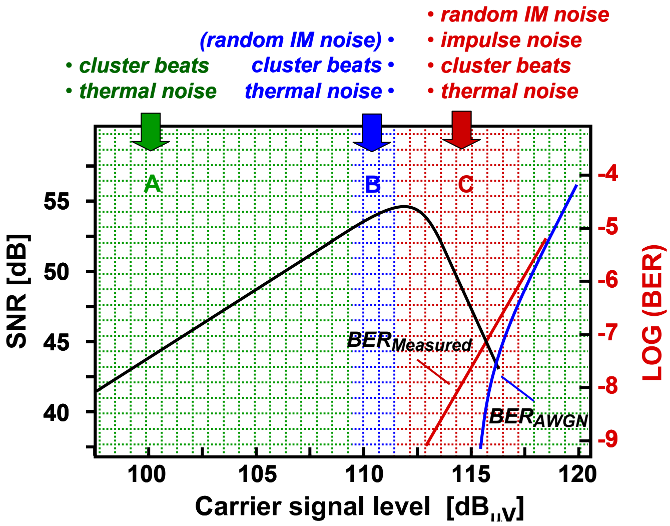


Figure 115: Schematic overview of the nature of non-linear distortion signals of an amplifier with a mixed analogue-digital load as a function of the carrier level of the digital signals.

In Figure 115 a schematic summary of the nature of the noise and non-linear distortion signal of an amplifier with a mixed analogue-digital load is presented. Irrespective of the carrier level, the noise and distortion signal encompasses the thermal noise of the amplifier and the narrow band cluster beats related with the intermodulation of the TV signals. For a higher signal level of the digital carriers, broadband random noise like distortion signals appear associated with the intermodulation of analogue with digital carrier and of digital carriers with digital carriers. In addition, impulse noise is generated for a high carrier level.

16.2 Degradation in case of digital and non-modulated carriers

According to the current understanding, signal degradation is generally assigned to the 2nd and 3rd order non-linear components. ReDeSign studies showed a rather different result; in case of digital signals the degradation is caused by the 4th and 5th order non-linear components instead of the 2nd and 3rd order. The 2nd and 3rd order contributions are generated like in the case of analogue signals; however, their contribution is masked by the thermal noise generated by the amplifier. In the following we will clarify this in detail using degradation data from an amplifier with a digital load and a CENELEC test load of 42 non-modulated carriers.

In case of the CENELEC load, only narrowband intermodulation products are generated. To assess the degradation we have measured the magnitude of the 2nd (CSO) and order 3rd (CTB) cluster beats for increasing carrier level using a spectrum analyser with a small bandwidth resolution of 50 kHz. Figure 117 shows the recorded carrier-to-intermodulation ratio (CINR) curves for a typical sample as measured for 3 frequencies. With a dedicated CENELEC test set up, we have measured the CSO and CTB performance figures as specified in IEC-60728 part 3 [10], paragraph 4.2. With the aid of these figures we have calculated the 2nd and 3rd order SNR curves for these frequencies. These simulated SNR curves are included in Figure 117.

A first look at the SNR curves for the CENELEC load shows that for high carrier levels the measured curves have a slope -1 and -2 for the 2nd and 3rd order distortions, respectively, as expected. Furthermore, the measured and simulated *CINR* curves match fairly well; the measured and simulated curves have a conform shape, but are shifted by some dB.

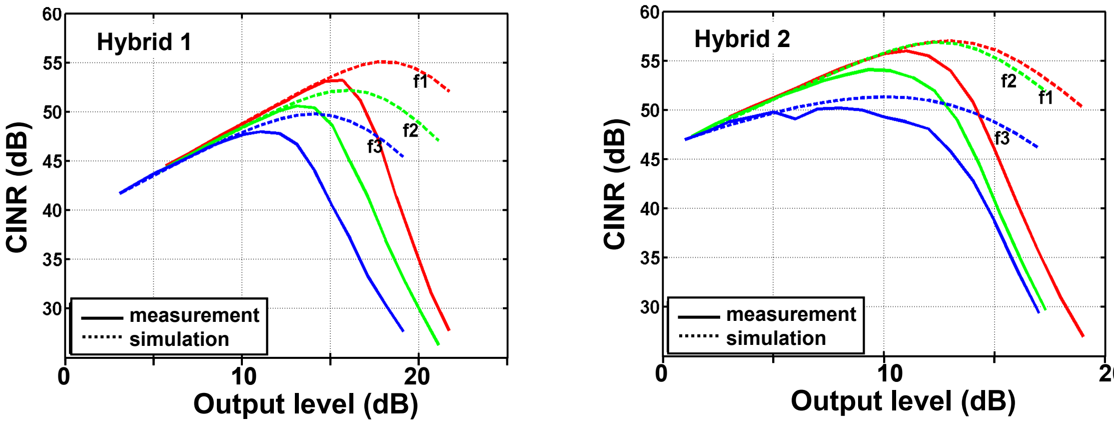


Figure 116: Measured and simulated SNR curves for 119 MHz (f1), 420 MHz (f2) and 855 MHz (f3) as obtained for a component with a composite load of 96 digital carriers. The bandwidth resolution was 8 MHz. The measured curves show a high carrier level asymptote with a slope -4. For the simulation a 2nd and 3rd order component model is used. The simulated and measured curves clearly are not congruent, showing that 2nd and 3rd order intermodulation does not dominate the SNR curves.

Next, we applied a composite load of 95 digital carriers to the same amplifiers and measured the SNR curves. The distortion signal was assessed in a free 8 MHz channel with a spectrum analyser with 8 MHz bandwidth resolution. The result for two amplifiers is shown in Figure 116. In addition, the figure shows the simulated SNR curves for the different frequencies. For these simulations we used a component model including 2nd and 3rd order non-linear behaviour. The 2nd and 3rd order model parameters were obtained from a standard CSO/CTB specification measurement using the before mentioned dedicated CENELEC test set-up.

The measured SNR curves of Figure 116 all reveal a very steep decline of the curve for high carrier levels. All curves approximately approach a decline with slope -4 which is associated with degradation by a 5th order non-linearity. Moreover, most curves show a rapid transition

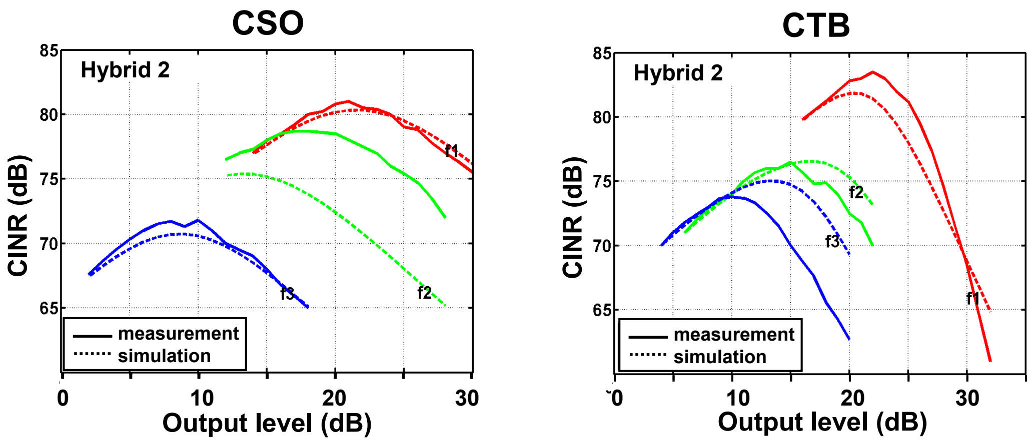


Figure 117: Measured and simulated CINR curves as obtained for a component with a (sloped) CENELEC load with 42 non-modulated carriers for 119 MHz (f1), 420 MHz (f2) and 855 MHz (f3). Measurement resolution was 50 KHz. The left panel shows the CINR for the CSO beats; the right panel shows the CINR for the CTB beats. For the simulation a 2nd and 3rd order component model is used. The measured curves show a high carrier level asymptote with a slope -1 and -2 for the CSO and CTB CINR curves respectively. The simulated and measured curves have congruent shapes, showing that indeed 2nd and 3rd order intermodulation dominates the CINR curves.

from slope +1 to a steep decline without any straight sections with slope -1 or -2 associated with degradation by 2nd or 3rd order behaviour. Comparison with the simulated curves shows that in all cases the 2nd or 3rd order component model yields a too optimistic *SNR* curve. Most remarkable is the lack of conformity of the shape of the measured and simulated curves.

The result of the comparison of the *CINR* and *SNR* curves for a CENELEC load and a digital load can be summarized as:

- In case of a CENELEC load the standard 2nd /3rd order model provides a qualitatively correct description of the degradation of the signals,
- In case of digital loads the standard 2nd /3rd order model does not provide a qualitatively satisfactory description, the results suggest that degradation is dominated by higher order non-linear contributions,
- In case of digital loads the standard 2nd /3rd order model provides a too low estimate of the distortion signal power, thus providing a too optimistic estimate of the *SNR* curve.

For completeness, one should note the SNR curves for digital loads currently are studied for reasons of specification of the components. As a rule these curves have an asymptote with slope -4 associated with 5th order intermodulation for high carrier levels in combination with a limited transition range from the part with slope +1 and without indications for intermediate regions with 2nd or 3rd order dominance. As such, the measured SNR curves of Figure 117 are representative for many amplifiers.

Considering the different behaviour of the *CINR* and *SNR* curves for a CENELEC and a digital load, the question concerns a straightforward and consistent explanation. The absence of a visible or a measurable 2nd and 3rd order degradation in case of digital carriers does not implicate that 2nd and 3rd order distortion products are generated; it only shows that these are not measured.

A simple and logical explanation is found when taking into account the thermal noise generated by the amplifier. The signal power of the thermal noise is proportional to the measurement bandwidth. In case of the CENELEC load, a 50 kHz measurement bandwidth was applied versus an 8 MHz measurement bandwidth in case of a digital load. This results in a $20\log_{10}(8000/50)$ or 23 dB difference of the noise level. Because of this, the measurement in case of the digital load is less sensitive and a significantly higher distortion signal level is needed to surpass the thermal noise level. If so, possibly the 4th and 5th order terms dominate the 2nd and 3rd order non-linear signal. In Figure 118 we have illustrated this explanation in a schematic manner.

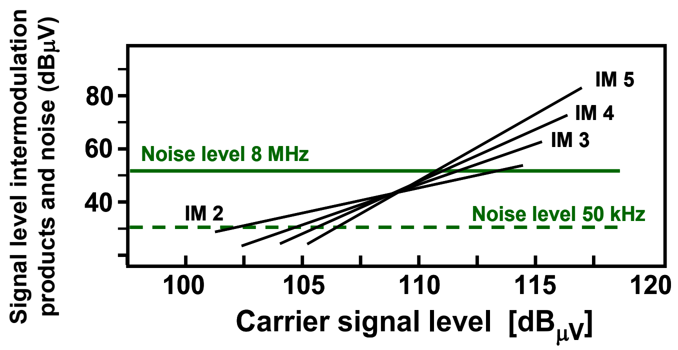


Figure 118: Schematic diagram of the signal level of the intermodulation products as a function of the carrier level. The noise levels are indicated for 50 kHz and with 8 MHz bandwidth resolution.

As an alternative way to explain the influence of the thermal noise, it is helpful to compare the cases of a composite load of digital (broadband) carriers and of non-modulated (narrowband) carriers, with the same number of carriers and the same average signal level. Thus both cases have a system load with a same composite signal power level, albeit composed of broadband signals and narrowband signals in the respective cases. Evidently, the signal power of the intermodulation products will be the same in both cases as well. However, the intermodulation products will be different in nature: the broadband load of digital carriers will generate broadband random noise evenly distributed in the frequency domain. In contrast, in case of the load of non-modulated carriers, narrowband cluster beats are generated. In case of the broadband signals the distortion signal power is completely smeared out over the full frequency range whereas in case of the non-modulated carriers the distortion signal is concentrated in a limited number of cluster beats with high spectral power density. Taking the thermal noise of the amplifier into consideration, the broadband intermodulation signal level is below or equal to this thermal noise level whereas the cluster beats peak well above the thermal noise level.

In summary we can conclude that the dominance of 4th and 5th order non-linear terms in the degradation of the signal appears a logical explanation of the observed shape of the SNR curve of a digital carrier. This explanation does not exclude the contribution of 2nd or 3rd order terms in the degradation of digital signals; in only provides a straightforward explanation for the fact that such 2nd and 3rd order degradation is not observed. This explanation does not exclude the possibility that, in case of amplifiers with a low noise figure and/or weak 4th and 5th order non-linear behaviour, the degradation of the digital signal is dominated by the 2nd and 3rd order terms.

Currently, there is no method available for specifying the 4th and 5th order non-linear of a component[6]. Therefore, one cannot make an appropriate simulation of the SNR curve. For network RF planning though, the capability of making a first rough guess would be valuable. Since we have no better method yet, it appears most appropriate to derive an estimate of the SNR curve for a digital load from the simulated SNR curve assuming a 2nd and 3rd order non-linear behaviour only. This 2nd / 3rd order SNR curve provides a too optimistic estimation of the real performance. From Figure 116, it can be read that the 2nd / 3rd order estimate is about 4 dB too optimistic; the maximum of the SNR curve is roughly shifted to a 4 dB higher carrier level. Since there is no better method yet, we propose the use of the above approach based on a 4 dB correction of the 2nd/3rd order SNR curve as a first rough guess of the performance for digital loads.

16.3 Signal quality parameters

The required quality of the signals at the system outlet for FM radio, analogue TV and DVB-C are specified in [4]. For all services, the signal quality must be compliant to the appropriate specifications.

[6] Within the ReDeSign project, TNO is currently studying the possibilities of the measurement of the 4th and 5th order terms and of simulation of the SNR curves with the aid of the UTOPIC RF planning tool for cable networks.

16.3.1 Analogue TV signal requirements

16.3.1.1 Analogue TV – The carrier-to-noise ratio (CNR_{aTV})

The measurement of the CNR_{aTV} signal is specified in IEC 60728-1 [4], paragraph 4.6 as the ratio between the carrier level and the noise level as measured in the 5 MHz video channel of the analogue signal. Thus the noise refers to the noise and distortion signal measured in a 5 MHz band. An indicative CNR_{aTV} requirement is given in IEC-60728 part 1 [4], paragraph 5.8. For PAL B, for example a 44 dB CNR value is required for a TV signal with a picture quality with mean opinion score (MOS) of 4.

16.3.1.2 Analogue TV – The carrier-to-intermodulation ratio ($CINR_{aTV}$)

The $CINR_{aTV}$ is defined as the ratio between the carrier signal and the weighted sum of the CSO and CTB cluster beats measured as specified in IEC-60728 part 1 [4], paragraph 4.5.3. In case of a periodic channel planning with an equidistant carrier distance of 7 or 8 MHz, the number of cluster beats in a 7 or 8 MHz analogue channel is limited. As a rule, these cluster beats are situated in a 30 kHz band; however, they have a higher spectral density than the thermal noise, as illustrated in Figure 113. The IEC-60728 part 1 [4], paragraph 5.9.3 provides an indicative $CINR_{aTV}$ requirement of 57 dB for a PAL signal. Like for the CNR_{aTV} ratio, this 57 dB $CINR_{aTV}$ ratio requirement refers to a MOS value of 4.

16.3.2 DVB-C signal requirements

16.3.2.1 DVB-C – The signal-to-noise ratio (SNR_{DVB-C})

Random noise causes a spreading of the constellation points of the constellation diagram, which may result in bit errors. The SNR for a DVB-C carrier (SNR_{DVB-C}) refers to the ratio between the signal power and the noise power as measured with a bandwidth resolution equal or less than the DVB-C channel width. In case of random noise, the minimum required SNR_{DVB-C} for quasi error free transmission, is specified in IEC 60728-1 [4], paragraph 5.8. For 64 and 256 QAM modulation respectively, an SNR_{DVB-C} of at least 26 and 32 dB is needed.

16.3.3 DVB-C2 signal requirements

For DVB-C2 currently the IEC 60728 part 1 [4] does not provide the signal quality requirements. However, DVB-C2 has been designed and optimized for the existing HFC networks. The simulation scenarios used for the system validations have been defined with the minimum network requirements as specified in the IEC 60728 part 1 [4] in mind. Therefore it is expected that DVB-C2 works properly in HFC networks which are in compliance with the IEC standard.

16.3.3.1 DVB-C2 – Signal-to-(random) noise ratio (SNR_{DVB-C2})

Currently, there is no information to specify a minimum signal-to-(random) noise ratio for DVB-C2 (SNR_{DVB-C2}) than the first system calculations [11]. These values may help as a starting point and have to be validated by forthcoming measurements data from real equipment and by performance figures from real network deployments. Table 16 list the minimum SNR_{DVB-C2} as obtained from simulations.

For completeness we have to mention that the (estimated) minimum SNR_{DVB-C2} requirement only applies to low digital carrier levels (DVB-C and DVB-C2), when the performance of the amplifier is limited by the thermal noise and when non-linear behaviour is negligible. For high digital carrier levels, driving the amplifier in the non-linear regime, impulse events will have an additional degrading impact on the performance.

Modulation	Code Rate (CR)	Noise Floor (dBµV)	SNR (dB)Fehler! extmarke nicht definiert.	Implementation margin (dB)	In-home margin (dB)	Minimum signal level (dBµV)
16-QAM	4/5	4	**10,7**	10	7	**31,7**
	9/10	4	**12,8**	10	7	**33,8**
64	2/3	4	**13,5**	10	7	**34,5**
	4/5	4	**16,1**	10	7	**37,1**
	9/10	4	**18,5**	10	7	**39,5**
256	3/4	4	**20,0**	11	7	**42,0**
	5/6	4	**22,0**	11	7	**44,0**
	9/10	4	**24,0**	11	7	**46,0**
1024	3/4	4	**24,8**	11	7	**46,8**
	5/6	4	**27,2**	11	7	**49,2**
	9/10	4	**29,5**	11	7	**51,5**
4096	5/6	4	**32,4**	12	7	**55,4**
	9/10	4	**35,0**	12	7	**58**

Table 16: Minimum $SNR_{DVB\text{-}C2}$ and minimum DVB-C2 signal level requirements

16.3.3.2 DVB-C2 – Signal level requirement

For obvious reasons, no requirement for the DVB-C2 signal level at the cable system outlet is currently included in the IEC-60728 part 1 specification [4]. Therefore a first estimation would be helpful as a guide for first network roll outs. Such a first estimate can be obtained from *i)* the sensitivity figure of a DVB-C2 receiver and by *ii)* allocating a signal margin for the in-home coaxial network between the system outlet and the DVB-C2 receiver. Obviously this signal margin concerns a business choice. In our consideration we have assumed a limited in-home network with a 2-port wall outlet (TV out and FM out) with a limited signal loss, a splitter to feed a secondary coaxial branch and of about 10-15m. In this scenario all losses will add up to about 7 dB.

Also the receiver sensitivity has to be estimated for the time being. In general the receiver sensitivity is composed of the thermal noise floor, the minimum signal-to-noise ratio for quasi error-free reception and the implementation loss and is obtained from a straightforward power budget calculation. The thermal noise floor of broadband cable technologies with an 8 MHz channel width is about 4 dBµV. Table 16 shows the required (estimated) $SNR_{DVB\text{-}C2}$ for quasi error free reception depends on the specific DVB-C2 modulation and protection schemes. Based on the implementation of the state-of-the-art DVB-C receivers, the implementation loss is estimated 11 dB for the DVB-C2 1024-QAM and lower QAM modulation modes and 12 dB for the 4096-QAM modulation mode. Adding all figures, as in Table 16 provides an estimation of the minimum DVB-C2 signal level.

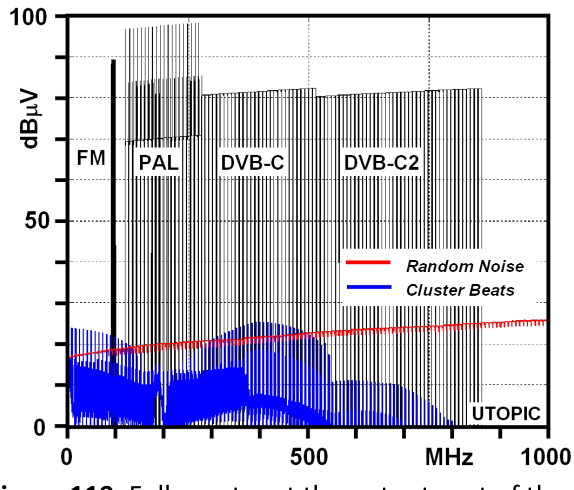
16.3.4 DVB-C2 RF signal planning

RF planning concerns the establishment of the most appropriate signal levels. Raising the signal levels improves the in-home reception of the signals and/or allows the application of a higher modulation scheme in case of digital carriers, but, for a too high composite signal level the signal quality will degrade due to the non-linear nature of the active components. Therefore, an operator has to make a trade-off between the signal level and the signal quality.

DVB-C2 will be introduced in an evolutionary manner. When launching DVB-C2 services, an operator will commence with a substantial but limited number of DVB-C2 carriers while preserving many, if not most, of the analogue TV and DVB-C services. For the existing analogue TV and DVB-C services, all operators have established specific signal levels that warrant a good in-home signal reception and signal quality. Therefore, when considering the roll out of DVB-C2, it appears most appropriate not to change these DVB-C and analogue TV signal levels. Thus the DVB-C2 planning challenge can be reformulated to the question of the appropriate DVB-C2 carrier level assuming an a-priori specified number of analogue TV, DVB-C and DVB-C2 carriers whereof the first two with a fixed signal level.

In Figure 119 we give an illustration of the impact of the DVB-C2 carrier level in such a planning scenario for a cascade with a mixed analogue digital load of 20 PAL carriers, 30 DVB-C carriers and 43 DVB-C2 carriers. For the PAL and DVB-C signal at the cable plant system outlet, respectively a level of 69 and 65 dBµV was assumed. This illustration is obtained from a simulation of a cascade of a number of amplifiers using a $2^{nd}/3^{rd}$ order component model[7]. The figure shows the spectrum at the output port of an amplifier of the cascade as measured with a spectrum analyzer with 300 kHz measurement resolution. More specific details of the simulations can be found in the Broadband issue of March 2010 [12]. The figure clearly shows that raising the DVB-C2 signal level results in a higher signal level of the random noise like non-linear distortion products, but that it does not affect the signal level of the narrowband distortion products (CSO/CTB cluster beats).

Using the above simulations, one can calculate the signal levels of the narrowband (CSO/CTB cluster beats) and broadband (random noise) distortion signals for all channels as a function of the DVB-C2 carrier level at the system outlet. Next the PAL carrier-to-noise ratio (CNR_{PAL}), the PAL carrier-to-intermodulation ratio ($CINR_{PAL}$), the DVB-C signal-to-noise ratio ($SNR_{DVB\text{-}C}$) and the DVB-C2 signal-to-noise ratio ($SNR_{DVB\text{-}C2}$) were calculated using the performance



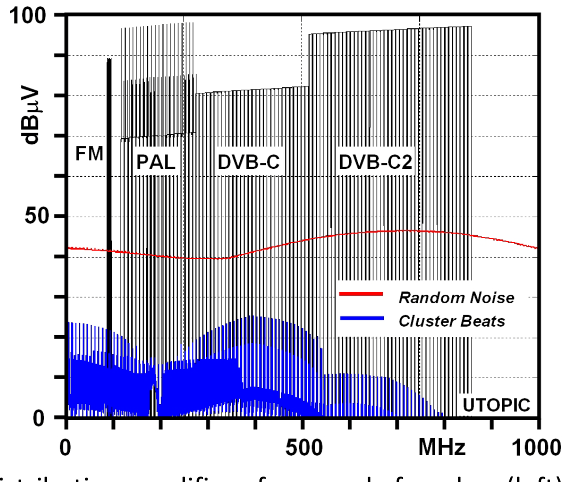


Figure 119: Full spectra at the output port of the distribution amplifier of a cascade for a low (left) and a high (right) DVB-C2 carrier level. The random noise (thermal noise and broadband intermodulation products) and the narrowband composite cluster beats are respectively shown in red and blue. The signal levels in dBµV refer to the level as measured with a spectrum analyzer with 300 kHz bandwidth resolution. Thus the real DVB-C and DVB-C2 levels are about 14 dB higher than shown.

[7] The model of Eq. 1 is used assuming zero 4^{th} and higher order coefficients, i.e. only CSO and CTB distortions are included.

definitions of section 16.3, and again as a function of the DVB-C2 carrier level. These values were calculated for all frequency channels, and the worst case value was used in the further analysis.

Such studies were performed for cascades which were specified in close cooperation with the operators. To be specific, the CSO and CTB and thermal noise specification figures of all components and the attenuation per cable segment were provided. The study included four cascades: a node and 2 amplifiers, a node and 4 amplifiers, a node and 5 amplifiers and a node and 15 amplifiers. In Figure 120 we show the (worst-case) $SNR_{DVB\text{-}C2}$ (top) and CNR_{PAL} (bottom) curves for all four cascades.

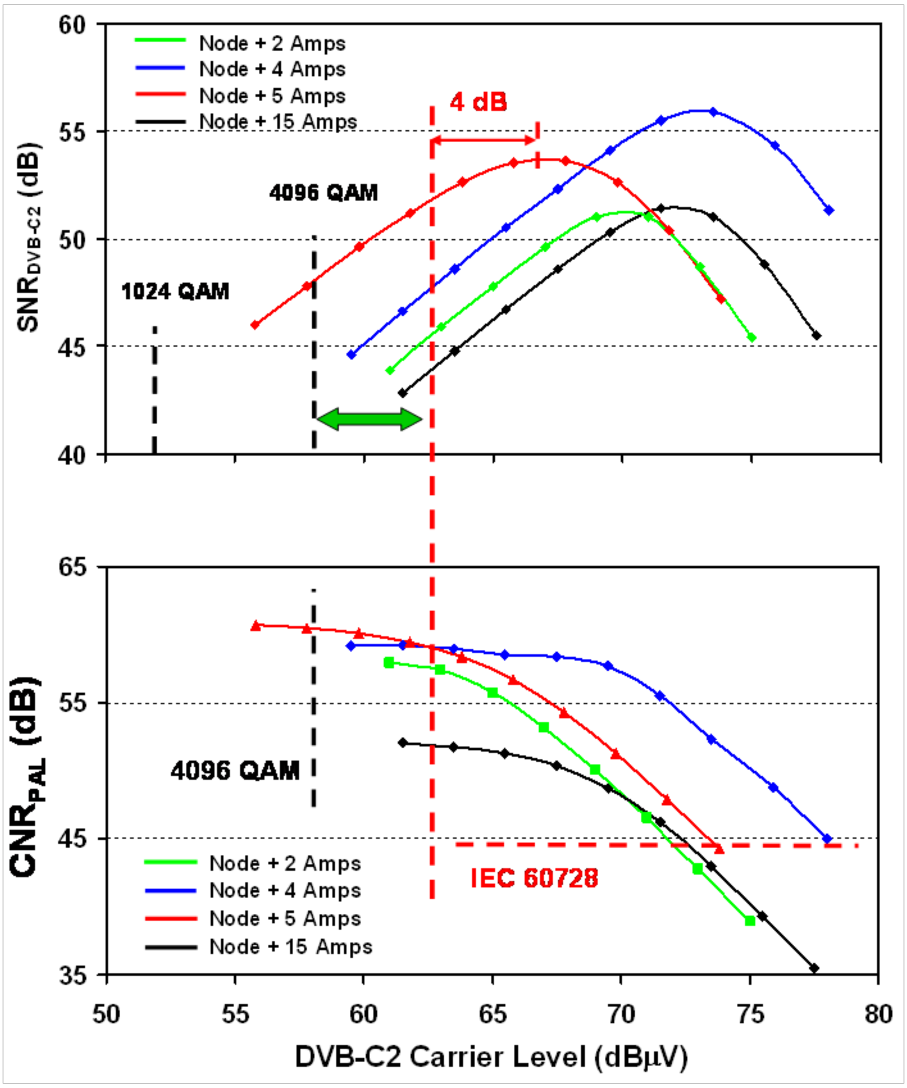


Figure 120: $SNR_{DVB\text{-}C2}$ and CNR_{PAL} versus the DVB-C2 carrier level for 4 cascades with a load of Figure 119 (20 analogue TV, 30 DVB-C and 43 DVB-C2)

The top panel of Figure 120 shows the (worst-case) $SNR_{DVB\text{-}C2}$ curve for a DVB-C2 signal level ranging from 55 up to 78 dBµV. In all cases the curves have a maximum associated with the onset of the generation of non-linear distortion products. The cascade of a node and 5 amplifiers (red curve), has of all SNR curves the most left curve. These calculations are based on a 2nd/3rd order component model, and therefore we should apply a 4 dB correction to find the maximum of the real $SNR_{DVB\text{-}C2}$ curve as explained in paragraph 16.2. From a planning viewpoint, the DVB-C2 carrier level associated with the maximum of the real SNR_{DVB-C2} curve marks the onset of impulse events and of the (steep) decline of the SNR_{DVB-C2}. For a robust signal delivery, an operator should be cautious not to apply a DVB-C2 signal level beyond this threshold. For the red curve (node and 5 amplifiers, the threshold DVB-C2 carrier level is indicated by the vertical dashed line.

In addition, the bottom window of Figure 120 shows the degradation of the (worst-case) CNR_{PAL}. As pointed out in section 16.3, this CNR_{PAL} parameter refers to the ratio between the analogue TV carrier level and the (random noise) distortion signal in the (broadband) video channel. Because of this, the degradation of CNR_{PAL} will occur in simultaneously with the degradation of the SNR_{DVB-C2}. The results of Figure 120 reflect this feature; the decline of CNR_{PAL} is aligned with the decline of SNR_{DVB-C2}. Moreover, the simulations show a safe margin for CNR_{PAL} with reference to the 44 dB requirement of the IEC/EN 60728 standard.

Like for the PAL signal, the DVB-C2 signal level was not changed. Because of this, the DVB-C signal-to-noise ratio (SNR_{DVB-C}) reveals a same behaviour as the CNR_{PAL}, as shown in the bottom panel of Figure 120, albeit shifted downward with about 4 dB because of the 4 dB back off of DVB-C with reference to the PAL carrier level. As argued before, and as shown in Figure 119, the carrier-to-narrowband interference ratio ($CINR_{PAL}$) is not affected by the DVB-C2 signal level since the magnitude of the narrowband CSO/CTB cluster beats is independent of the DVB-C2 signal level.

The vertical black dashed lines indicate the minimum DVB-C2 signal level needed for the proper reception of a DVB-C2 1024 QAM and 4096 QAM signal as taken from Table 16. Comparison of this minimum DVB-C2 carrier level and the maximum DVB-C2 carrier level shows whether or not a specific modulation scheme can be deployed. In all four cascades shown in Figure 120, deployment of the 4096-QAM mode appears feasible.

16.4 Final Remarks

In this chapter we have analyzed the degradation of the analogue and digital signals when deploying DVB-C2 services. In case of a mixed analogue digital network load, the non-linear distortion products consist of *i)* narrowband cluster beats (CSO/CTB beats) and *ii)* broadband random noise and *iii)* impulse events.

The narrowband cluster beats are associated with the intermodulation of the analogue TV signals. As such, the narrowband cluster beats are not affected by the signal level of the digital carriers, either DVB-C or DVB-C2; only the number of the analogue signals and the analogue carrier level define the magnitude of the cluster beats. Thus reducing the number of analogue carriers results in an improved carrier-to-narrowband interference ratio (CIR_{PAL}).

The broadband random noise is generated by the intermodulation of digital carriers with analogue carriers and digital carriers with digital carriers. This random noise associated with the non-linear nature of active components is of the same nature as the thermal noise. Because of this, the broadband random noise generated by intermodulation becomes notable at a sufficient composite digital carrier level only. This degradation is reflected in a (steep) decline of the signal-to-noise ratio for a carrier level above a specific threshold. Because of the broadband nature of the digital carriers, the degradation is generally dominated by 4^{th} and 5^{th} order non-linear behaviour of the amplifiers and not by 2^{nd} and 3^{rd} order terms.

Analogue TV services are degraded by broadband random noise as well. Because of the 5 MHz bandwidth of the video channel, the degradation of the CNR_{PAL} is dominated by the 4^{th} and 5^{th} order non-linear terms, like the degradation of the digital carriers. This degradation is quantitatively related with the degradation of the digital carriers, and takes off at about the same digital carrier level where the degradation of the digital carrier begins. Therefore, the analogue TV signal is sufficiently protected as long as the degradation of the digital carriers is avoided.

In summary we conclude that the existing understanding of the non-linear degradation assuming the dominance of 2^{nd} and 3^{rd} order effects fails in providing a proper description and understanding of the degradation of broadband signals, digital carriers as well as the analogue TV video signal. 4^{th} And 5^{th} order behaviour has to be taken into consideration.

From a practical RF planning viewpoint though, we can conclude that the signal quality is guaranteed as long as the maximum of the $SNR_{DVB\text{-}C2}$ curve is not surpassed; neither the $CINR_{aTV}$ nor CNR_{aTV} will degrade as long as the DVB-C2 carrier level remains beneath this threshold.

Finally it should be mentioned that the author is indebted to VECTOR (Poland) and Kathrein (Germany) for providing the technical facilities and support for the study.

17 Head-end Architectures

Headends are playing an important role within the transmission chain of TV signals as they are a bridge from contribution (e.g. by satellite) to distribution (e.g. via cable TV networks). CATV headends contain processes like reception, demodulation, decoding, processing, encoding and re-modulation of a variety of radio, TV and data signals. The focus within this report is on digital processing and modulation in accordance with the DVB standards. Beside the well known traditional processing of analogue radio and TV-Signals digital transmission equipment has been installed since 1995 within CATV headends. The appropriate standard for modulation was and still is DVB-C. It has to be pointed out that already DVB-C allowed a broad flexibility in terms of QAM constellations (from 16 QAM up to 256 QAM single carrier modulation), channel bandwidth (e.g. 6, 7, 8 or 12 MHz), different symbol rates and so on. Most of the modulation equipment from all manufacturers had this flexibility while the corresponding parts within the transmission chain -the Set-Top-Boxes- didn't have that flexibility also many constraints of network planning obligations have been the reason that today most of DVB-C signals are transmitted either as 64-QAM@8 MHz or 256-QAM@8 MHz signals only. Many CATV networks today are already full to capacity for the transport of digital signals and furthermore new tools and performance improvements are needed to address both private and business customers, particularly with IP-based content. Therefore the new DVB-C2 standard is an important step to fulfil the new requirements and is also a challenge for the industry to create and develop suitable modulation equipment which can be installed in existing network infrastructure.

17.1 End-to-end transmission chain

Since the early days of CableTV transmission the general end-to-end chain with its building blocks hasn't changed in general while many changes happened in detail. Figure 121 is showing the principle as it is still valid, starting within the studio, ending within subscriber premises. The focus here is of course the CableTV network with the processing part – the headend.

If we look into the details of the processing part of the transmission chain we find a variety of digital standards which have to be taken into account by realization of headend technology. Beside MPEG 2 and MPEG 4 the most important standards are DVB standards. The standards of the first generation DVB (DVB-S, -C and -T) have been adopted since years and the relevant headend equipment is state of the art since a long time.

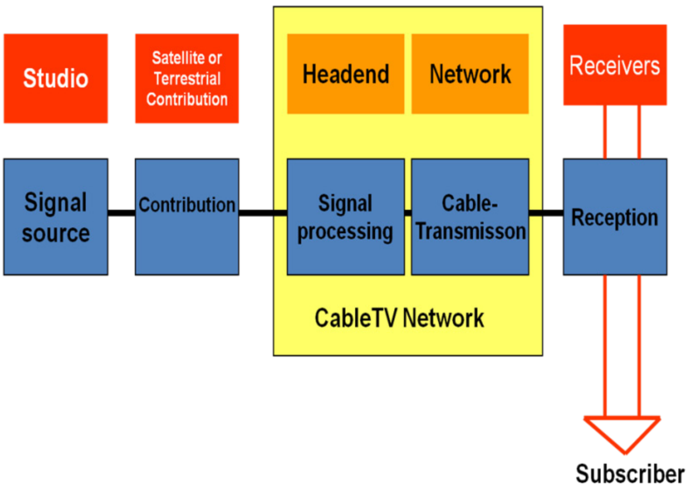


Figure 121: End-to-end transmission chain for TV-signals

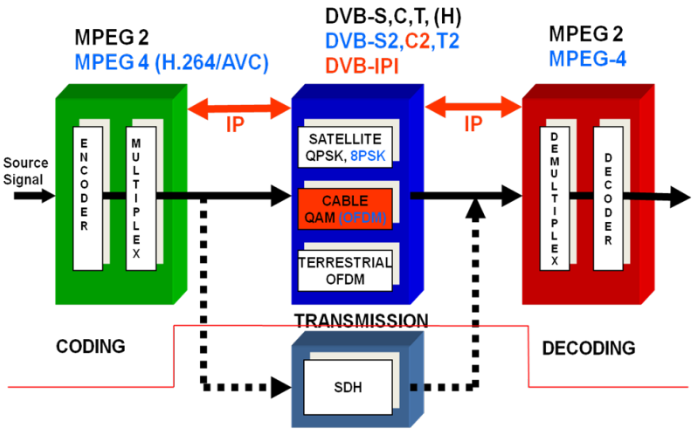


Figure 122: Standards for Digital TV – the tool box

In parallel to the introduction of the second generation of DVB standards (which happens now) it is important to point out that the IP technology (e.g. DVB-IPI) has an enormous impact on today's headend technology. So there is now a broad variety available within the tool box for standards for digital TV. Figure 122 is showing the scenario.

17.2 Headend structures: DVB technology – on the way to IP-structures

Within a typical headend (Figure 123) signals from different sources are received and processed. The physical link between the building blocks of existing headend infrastructure is in many cases the ASI-Interface. This has to be taken into account by planning the introduction of DVB-C2 modulation equipment into an existing environment. It is also a typical feature for those headends that content gets bundled locally by using multiplexers.

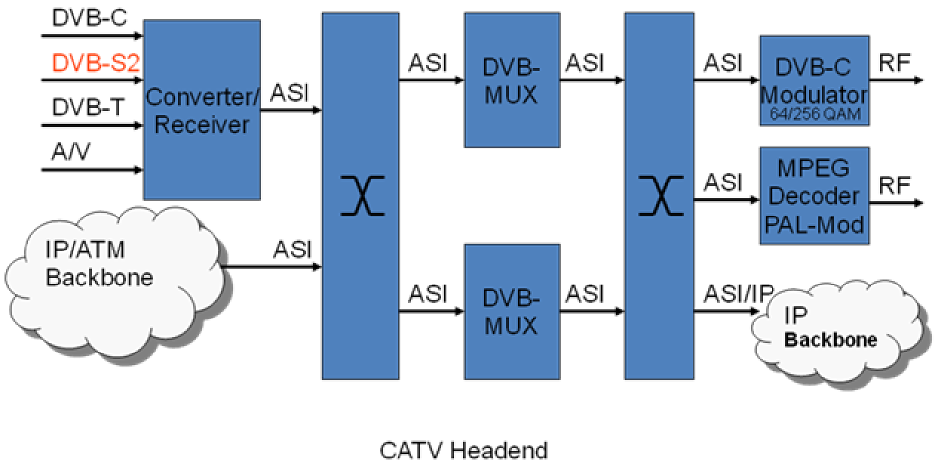


Figure 123: Traditional headend architecture

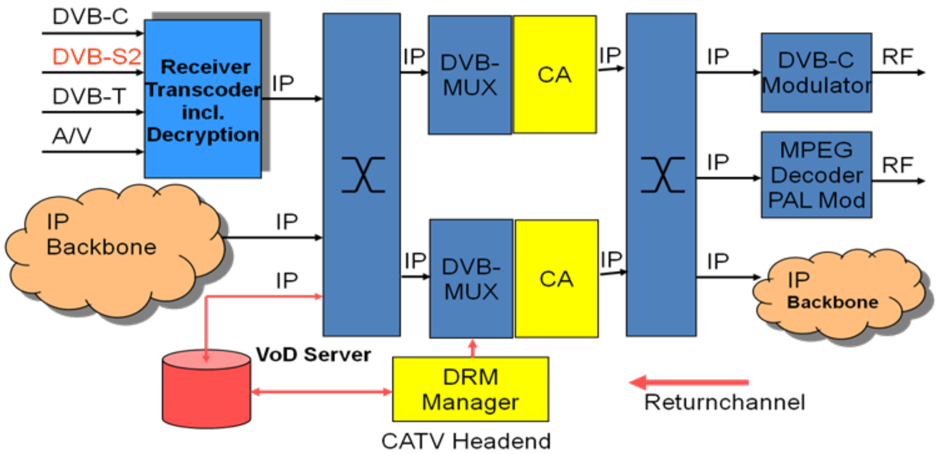


Figure 124: Enhanced headend architecture

17.3 Improvements of network- and headend infrastructure

CATV networks today are offering much more than just the distribution of TV services. The result is an improvement of network and headend infrastructure. The migration of broadcast and interactive services to triple play networks is taking place. At the same time we find so called master/sub-headend concepts using IP backbones which are replacing more and more the traditional CATV headends. Consequently new modulation schemes like DVB-C2 can help to increase the capacity in terms of transported data.

Within a master/sub-headend concept the reception and central processing is done at a central location (master headend) where for instance all outdoor equipment (antennas, satellite dishes etc.) is installed. Finally all signals are converted into IP signals (GigE) and get distributed via IP backbones to de-central locations (sub-headends, hubs or edges). In order to provide a redundancy in case of a total loss there is very often a second master headend connected to the backbone.

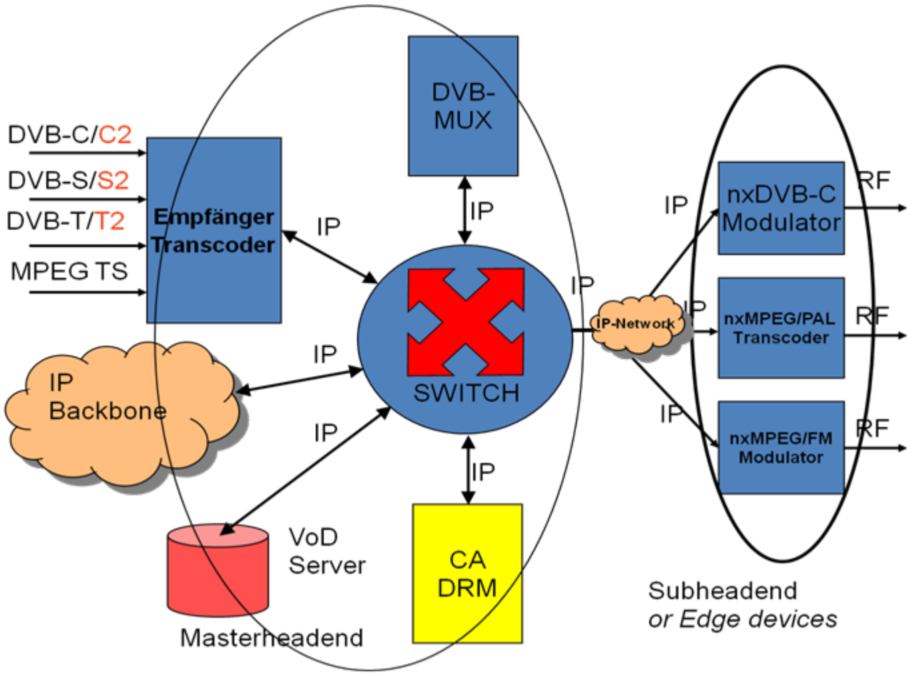


Figure 125: Master/Sub-headend architecture – edge devices

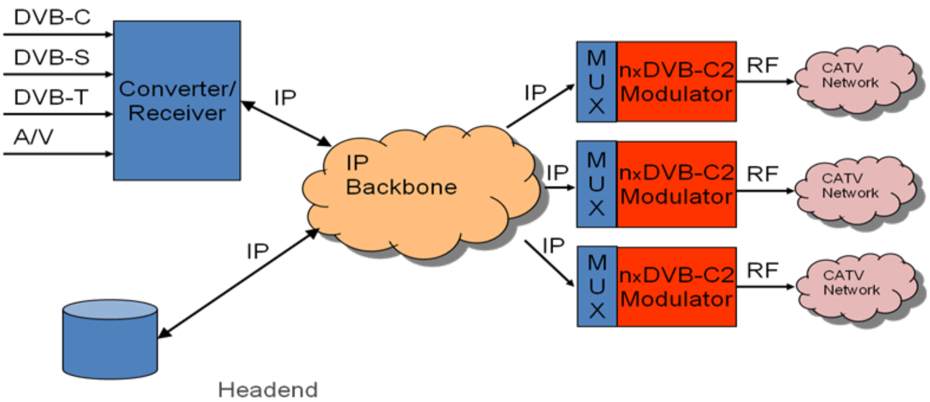


Figure 126: IP headend architecture using Edge QAMs

At the hub locations (sub-headends) the so called edge devices are connected to the backbone. These edge devices convert the relevant IP signals into “normal” radio and TV signals which can be fed via CATV networks to the subscriber's premises. For the time being there will be the trans-coding to analogue PAL signals and FM radio signal in parallel to DVB-C and DVB-C2 signals as shown in Figure 125. In the future, it is expected that it will be a complete digital scenario as shown in Figure 126.

17.4 Realization of modulator technology

As described in detail in the previous chapters one important difference between DVB-C and DVB-C2 is the modulation scheme. While DVB-C uses a single carrier modulation DVB-C2 benefits from a multi carrier modulation. This has to be taken into account by designing of filter and amplifier blocks. Another important difference which has influence on the modulator design is the input interface for multiple transport stream and generic stream encapsulation (GSE) for DVB-C2 modulators (see Figure 127).

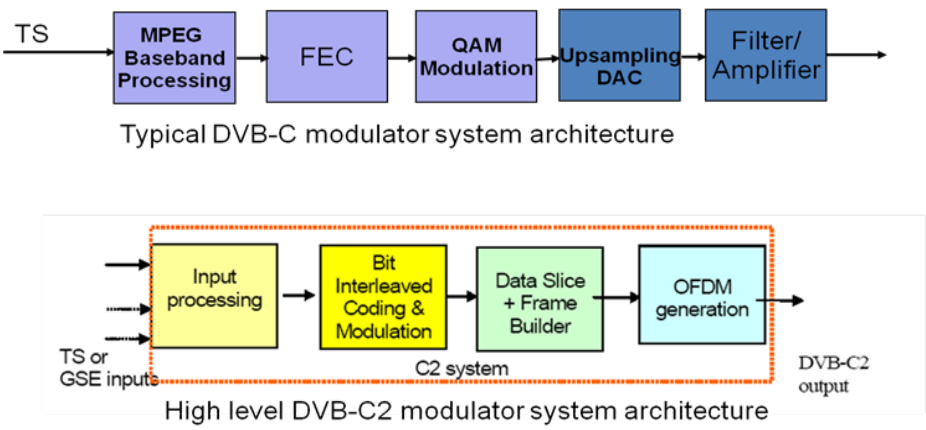


Figure 127: DVB-C modulators vs. DVB-C2 Modulators

A brief comparison of modes and features between DVB-C and DVB-C2 modulators is shown in Table 17.

	DVB-C	DVB-C2
Input Interface	Single Transport Stream (TS)	Multiple Transport Stream and Generic Stream Encapsulation (GSE)
Modes	Constant Coding & Modulation	Variable Coding & Modulation and Adaptive Coding & Modulation
FEC	Reed Solomon (RS)	LDPC + BCH
Interleaving	Bit-Interleaving	Bit- Time- and Frequency-Interleaving
Modulation	Single Carrier QAM	COFDM
Pilots	Not Applicable	Scattered and Continual Pilots
Guard Interval	Not Applicable	1/64 or 1/128
Modulation Schemes	16- to 256-QAM	16- to 4096-QAM

Table 17: Table comparing available modes and features in DVB-C and DVB-C2

Many developments of DVB-C modulators were based on available IC (Integrated Circuit) or ASIC (Application Specific Integration Circuit) technologies. The development steps for DVB-C2 modulators will be carried out first of all as verifications based on computer simulations. Hardware prototypes and components for volume production will be most probably developed on FPGA (Field Programmable Gate Array) technology.

Due to the fact that there are significant differences between the modulation schemes of DVB-C and DVB-C2 there is no chance to upgrade already installed DVB-C modulators to

DVB-C2 modulators. There are many installations where the DVB-C modulator is one portion of a complete trans-modulator (e.g. from DVB-S/S2 to DVB-C). In these cases we have to distinguish between the compact product approach and the modular product approach. As shown in Figure 128 and Figure 129 in case of a modular system only the modulator has to be replaced by a new DVB-C2 modulator. The other parts (e.g. DVB-S/S2 receivers) can remain. Compact trans-modulators have to be replaced where appropriate.

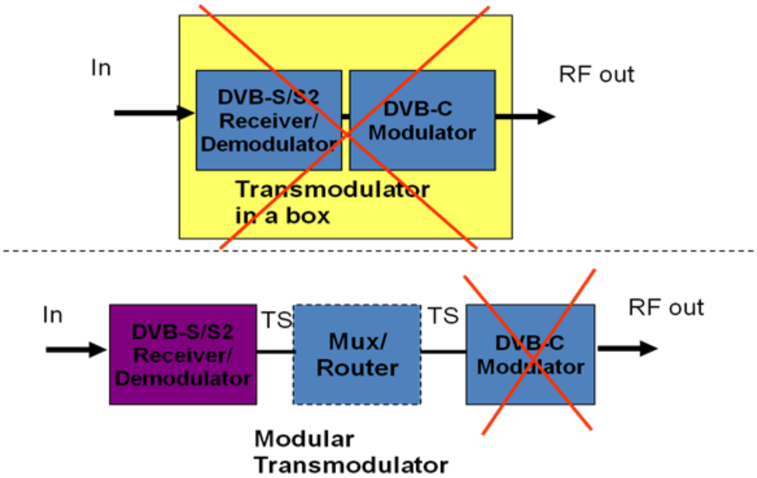


Figure 128: Impact on existing trans-modulators (DVB-C)

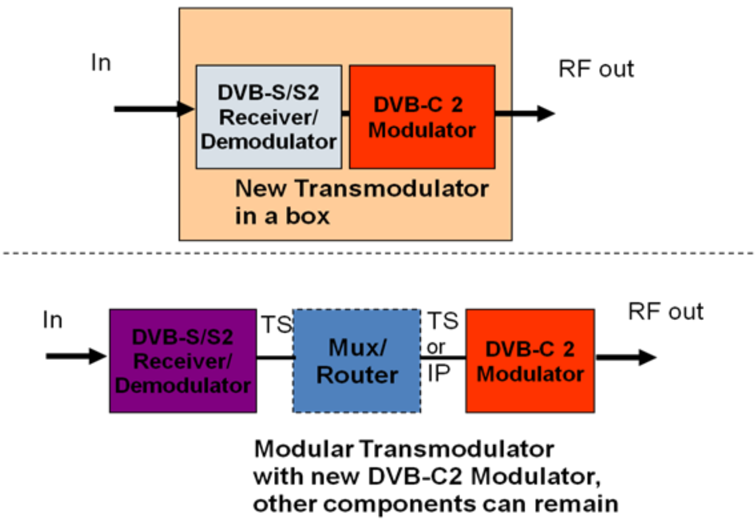


Figure 129: Impact on existing trans-modulators (with DVB-C2)

The DVB-C 2 standard provides an enormous advantage for new installations of modulation equipment. The PLP concept (Physical Layer Pipes) which is described in chapter 6 allows the design of modulation equipment with multiple inputs fed also directly from IP sources. A principle high level block diagram is given in Figure 130. By using this capability it will be possible to reduce the number of external multiplexers.

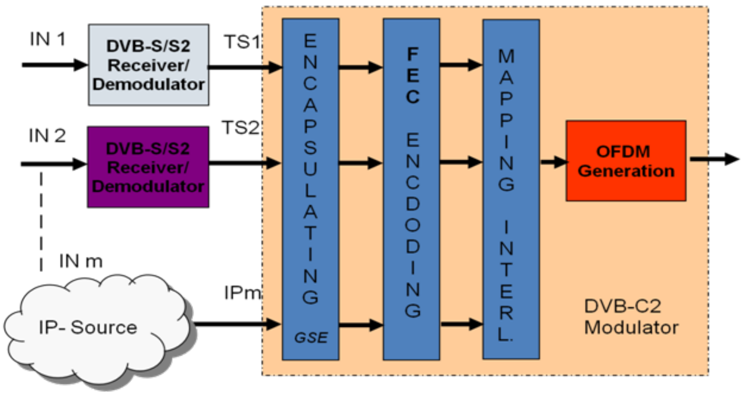


Figure 130: Enhanced DVB-C2 modulator solutions: PLP concept can replace re-multiplexing.

18 Implementation Scenarios

Current cable networks will run out of bandwidth since services like HDTV and HSI with 120 Mbps and more will increase significantly over the next few years. A new transmission technology is needed to resolve this capacity demands. Satellite and Terrestrial environments have already launched new standards to fulfil future service requirements. Cable needs to follow to compete also with incumbent Telecommunication providers (Telcos). Melting down analogue TV will provide the best chance to take over the frequencies used for the provision of this service with the new and efficient transmission technology DVB-C2. This chapter gives some considerations on how DVB-C2 could be introduced in existing MSO service portfolios.

18.1 Chanel line-up example

The following Figure 131 shows an example of a channel line-up used in an existing modern cable network. It explains which services are used at which frequencies for upstream and downstream frequencies up to 862 MHz.

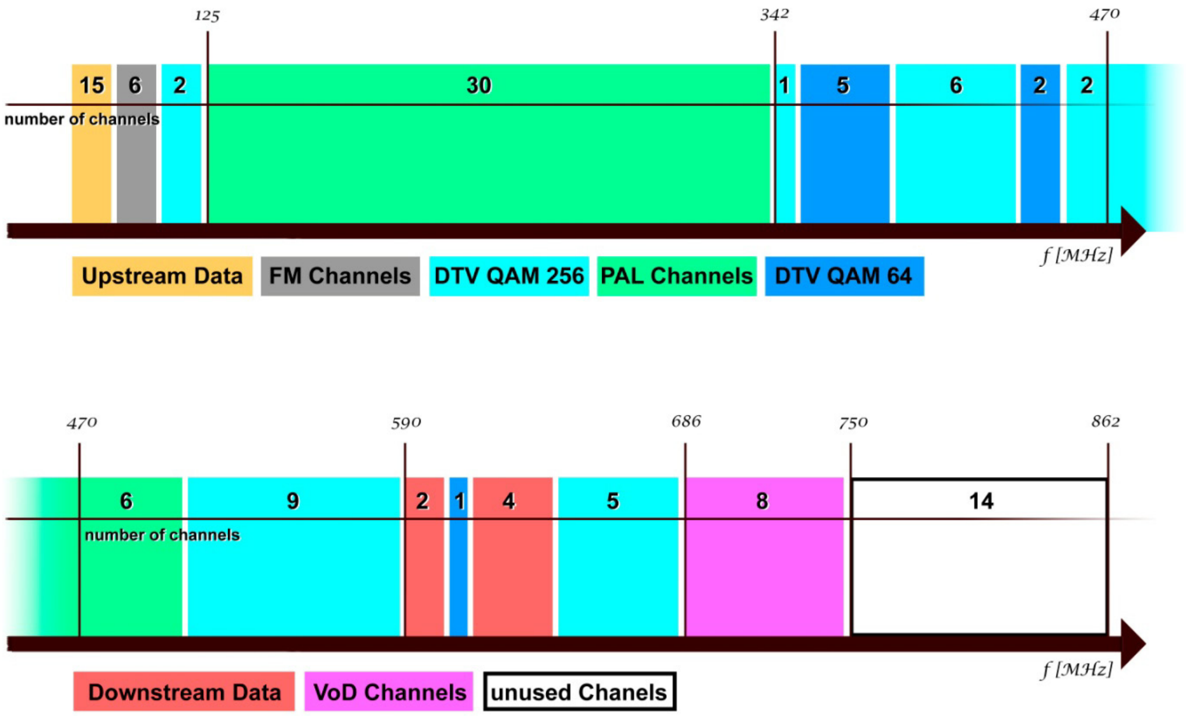


Figure 131: Example of channel line-up

In this example 30 channels are used for analogue TV providing a large bandwidth which could be used much more efficient when switching to DVB-C2. Since a 7 MHz channel grid is used in a wide range of this frequency band between 125 MHz and 342 MHz, a simple switch to 8 MHz wide digital signals would not be easily possible. MSOs have made some experience with the conversion of the mixed 7-8-MHz grid used by analogue TV signals to a uniform 8-MHz-grid required for DVB-C and EuroDOCSIS. DVB-C2, however, is based on OFDM and – as explained in various chapters above – has a very frequency agile bandwidth characteristic. It therefore supports the conversion of the channel grids very efficiently and, thus, can be used for DTV in a very flexible manner. If, in a first step, two channels should be freed, the available spectrum can be fully occupied by a DVB-C2 signal supporting a bandwidth of 14 MHz.

The missing link in this example is the broad offer of broadcasted HDTV services. In addition it is assumed that DOCSIS downstream signals and VOD services will heavily increase in the coming years. The 14 channels between 750 MHz and 862 MHz, which are available in the example considered here, provide noisy channel conditions which do not allow the utilization of DVB-C with a 256-QAM. In fact only 64-QAM can be used for DVB-C and DOCSIS. According to the simulation results presented in sub-chapter 15.4, DVB-C2, however, would support the provision of a much larger bit rate using 256-QAM with a Code Rate of 9/10 or even a 1024-QAM with a Code Rate of 3/4 (compare Figure 110).

18.2 Services offered

As mentioned in various chapters above, a cable network provides different categories of services. Today the dominant ones are broadcasting services in the very most cases:

- Analogue TV / FM (PAL, SECAM)
- Digital TV (SD, HD) / dRadio (DVB-C)
- High Speed Internet (DOCSIS 2.0)

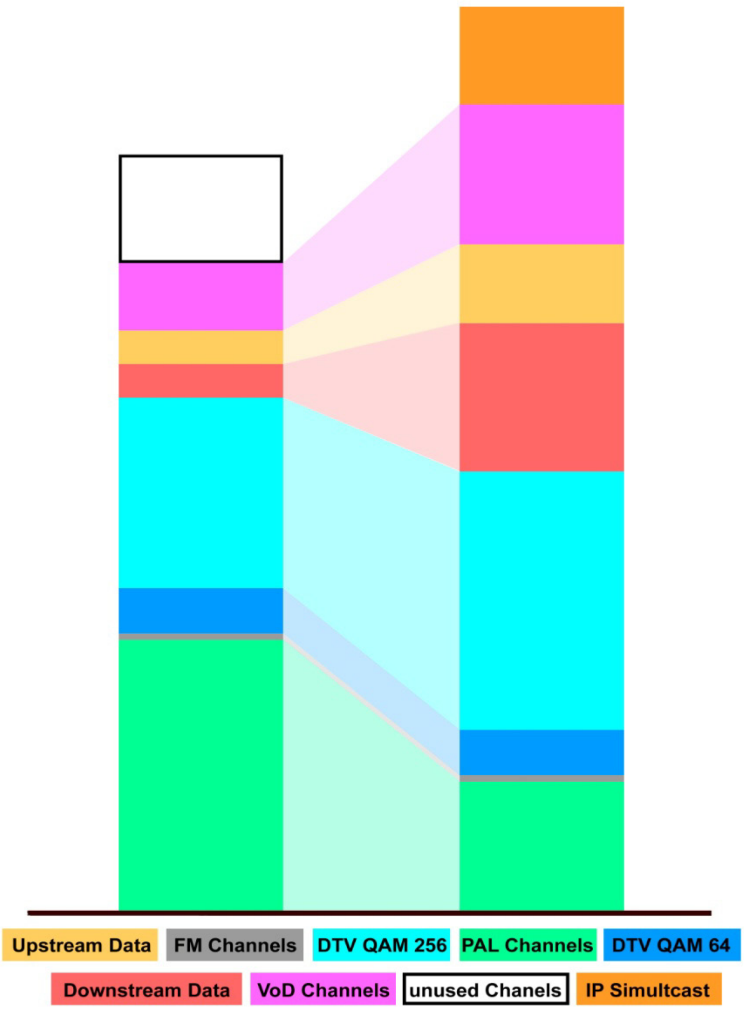


Figure 132: Example of bandwidth increase caused by take-up of new services

However, sophisticated services are emerging increasingly or are being introduced in short term:

- Video on Demand
- Switched Digital Video
- Video over DOCSIS (to support full IP gateways)
- High Speed Internet with even higher speeds of 120 Mbps and beyond (through channel bonding in combination with DOCSIS 3.0)

In particular the increasing offers of HSI and HD content demand additional bandwidth. An example of the expected change in bandwidths allocation by MSO services is depicted in Figure 132. HSI downstream is indicated as data downstream and HD content is part of the DTV 256-QAM and VOD services. It seems to be obvious that the requirement for more transmission capacity can be fulfilled by introducing a physical layer transmission system being more efficient and flexible than DVB-C and DOCSIS developed 15 years ago.

18.3 DOCSIS using C2 Technologies

DVB-C2 also provides an efficient way of transmitting IP traffic with Generic Stream Encapsulation (GSE) which can be used instead of the widely introduced Transport Steam. It is therefore worthwhile considering to make use of big pipes provided by the new physical layer and the gain of bandwidth not only for DTV and other broadcast and push services, but also for IP downstream applications. Thus, DVB-C2 seems to be a very interesting candidate to be integrated in DOCSIS and in the European DOCSIS version, respectively. The idea of an integrated system approach which can be used for all kinds of data such as TV, HD, and HSI is still an important aspect for MSOs to consider. Such integration would entail various advantages; the reduction of the CMTS platform complexity and the replacement of channel bonding through a flexible channel bandwidth management at physical layer are mentioned by way of examples. An integration DVB-C2 / DOCSIS full-service platform receives particular interest when moving to full IP gateway solutions providing video over DOCSIS while using DVB-C2 downstreams. Such a solution will among others allow transporting all TV and radio services in a traditional as well as in a combined manner with an integrated service portfolio for TV, radio, HSI, and telephony etc. on the same platform and in the bandwidths.

18.4 Implementation Scenario

The introduction of DVB-C2 will by no means take place in a single step. It is expected that a rather smooth transition will happen where individual service cluster are migrated from the traditional DVB-C/DOCSIS platform to the state of the art platform based on a modernized DOCSIS and DVB-C2. Depending on the services already launched, it will be possible to support legacy STBs over long time on DVB-C. The same applies to the different generations of DOCSIS.

Using the analogue spectrum for the implementation of DVB-C2, the increase of available bandwidth and the broader transmission capacity of the network would allow postponing expensive network segmentation by node splitting into the future. Provided that DVB-C2 would be integrated in DOCSIS, all services could be moved to a DVB-C2 spectrum over time which would significantly reduce the segmentation issue.

This final goal could be reached if all CPEs (STBs, cable modems (CMs), Gateways) used in the network are in compliance with DVB-C2. The wide introduction of the equipment is a cost issue which requires a detailed analysis of the situation of each individual MSO. In particular Number of legacy CPE in terms of STBs and CMs compliant with DVB-C-and traditional DOCSIS is of interests as well as the age of the equipment and in how far it is depreciated.

Figure 133 below displays an example of a STB roadmap fostering the strategy of full service provision over IP in the future. Starting from DVB-C CPEs, the final goal is to have devices deployed which are based on IP only. Gateways are expected to support all services (e.g. Internet, telephony, TV) and allows for an IP based content distribution and content sharing in the home. Eventually simple and thin IP clients are expected to replace DVB STBs. Portal software will allow for consuming content on PCs.

The migration from DVB based to IP based services requires the reduction of legacy equipment which can be achieved by different means such as: first of all the shipment of aged devices needs to be stopped as soon as the new DVB-C2 CPEs are available in order to freeze the legacy DVB-C STB population. Old devices such as SD zapper boxes and DOCSIS 1.0 and 2.0 CMs which have the lowest penetration are the preferred candidates. The refurbishment of headend and CPE equipment needs to be reduced to a number required to maintain the legacy service on a necessary level. For instance, HD-DVRs launched with DVB-C downstream but equipped with adequate Ethernet ports can still be used in a full IP environment by serving the device with content provided via a DOCSIS Ethernet port.

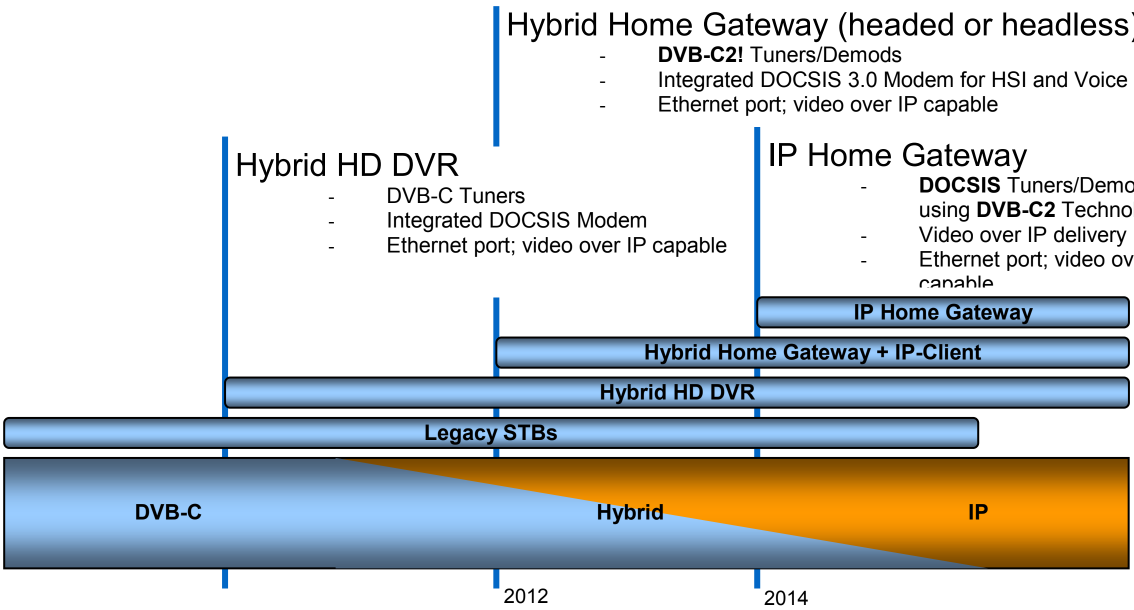


Figure 133: Example of set-top box roadmap

Furthermore marketing campaigns needs to be launched to convince customers about the advantages of the new system thus moving to a new product.

First services which will be implemented on DVB-C2 are interactive TV services such as VOD using the return path via DOCSIS. Such services are capable of working on the traditional as well as on the new platforms in parallel without the need to increase the bandwidth. The headend is aware of the kind of CPE and sends the demanded service either via the DVB-C/DOCSIS of via DVB-C2. In contrast the introduction of DVB-C2 for broadcast services would indeed increase the bandwidth since simulcast transmission of a single service via both platforms DVB-C and DVB-C2 would be necessary. This redundancy in fact could be prevented in case of unicast services.

In the following a brief overview on implementation steps are given to prepare for a DVB-C2 rollout. In a first step, STBs and hybrid gateways with DVB-C2 tuner need to be deployed taking account backwards compatibility to DVB-C of the devices. The capability of the new boxed to receive not only DVB-C2 but also DVB-C is of utmost importance. In parallel, DVB-C2 compliant EdgeQAMs need to be deployed. Transmission capacity made available for a DVB-C2 rollout allows for the introduction of a VOD services which could be operated in broadcast mode at the beginning. Next generation EdgeQAMs should be equipped with appropriate processing power and programmable elements to have sufficient flexibility for the support of different use cases. It must be possible to fully migrate to DVB-C2 by firmware updates on deployed DVB-C2 capable hardware. In a parallel effort it would be worthwhile to develop an advanced DOCSIS standard using DVB-C2 technologies for downstream. Various existing interfaces of DOCSIS would need to be updated for this purpose. Most efficient technologies should be selected such as GSE, COFDM, ACM, 4K-QAM etc.

A second implementation step in the rollout scenario could be to start using DVB-C2 to provide in parallel to DVB-C more capacity for additional services and for on-demand services in particular. The deployment of DVB-C2 capable STBs and gateways need to be fostered while slowly decreasing the number of DVB-C compliant STBs from older VDD rollout. These efforts will result in a reduction of bandwidth assigned to service provided via DVB-C in general. The number of frequencies used for the different technologies may vary between different network clusters and service areas, respectively, and depend on the number of STBs deployed and the number of services consumed. It therefore will be necessary to manage the allocation of capacities between DVB-C and DVB-C2 based transmissions per area; probably steps of 2 to 4 frequency bands are necessary for this purpose. The availability of frequencies is important for the introduction of a new technology, thus, the reduction of frequencies used for analogue TV distribution is much desired. As explained above, a 7 MHz channel grid is used in the lower frequency band of the cable downstream spectrum which prevents a simple utilization of signals with 8 MHz bandwidth. However, the frequency agility of DVB-C2 allows an efficient combination of, for instance, two 7 Hz channels and thus the injection of DVB-C2 signals of 14 MHZ bandwidth each. The lower downstream frequency band has a particular high value since it provides more robustness and thus a better prerequisite for the introduction of higher modulation schemes. The implementation step has a timing which heavily dependent on the availability of DVB-C2 STBs and headend equipment as well as on the achievements to melt down analogue TV within the given market environment.

Last but not least the third implementation step focuses on the use of DVB-C2 for HSI and the move to Video over IP. This could be achieved by introducing DVB-C2 for DOCSIS downstreams as described above while freezing the rollout of DOCSIS 3.0 cable modems. The introduction of M-CMTS available for the connection to DVB-C2 compliant EdgeQAMs or DVB-C2 fully integrated CMTS would be continued in parallel to the deployment of IP gateways and IP client STBs. The provision of IP multicast and on-demand video services over DOCSIS could be implemented step by step and region by region. During this time earlier deployed STBs will be supported via appropriate numbers of DVB-C/C2 frequencies. The number of the required frequencies differs from cluster to clusters depending on bandwidth demands of customers with 1st generation VOD boxes. Frequencies are made available to serve the advanced DOCSIS platforms using DVB-C2 for HSI downstream transmission. Older DOCSIS versions being still in operation will have to be continuously supported for a certain time. This will lead to a higher degree of spectrum segmentation for a limited period. In order to avoid further spectrum segmentation in the future, it is recommended not to launch an advanced DOCSIS platform which would not make use of the benefits of DVB-C2. The timing of this implementation step dependent on the availability of appropriate numbers of frequencies to cover IP simulcast of broadcasted services (preferably out of the today's analogue TV spectrum) as well as the availability of the advanced DOCSIS platform incorporating DVB-C2.

19 Annex

19.1 Responsibilities and Acknowledgements

The individual chapters of this book have been prepared by the authors mentioned. The responsibility of the content of the chapters is entirely with the respective authors.

The authors thank their companies for supporting the preparation of the book by making time and man-power available. Furthermore the authors thank the DVB Project for providing the facilities necessary to prepare the book. The cooperation among the members of the DVB experts groups during the development of the DVB-C2 standard and the accompanying documentation including this book was excellent. Thanks also to the members of the ReDeSign project who joined the DVB work through a liaison agreement signed by the two organizations. The ReDeSign work as well as the publication of this book was co-funded by the European Commission. Many thanks for this sponsoring. Last but not least, ANGA was instrumental in organizing an Implementers' Workshop which was the priming for gathering all the content eventually leading to this book.

19.2 Bibliographies of Authors

19.2.1 DR. PETER SIEBERT

Peter received his M.Sc. in 1984 and his Ph.D. in 1989 in physics from the J.W. Goethe University in Frankfurt, Germany. He has been a member of the IEEE for 23 years and a Senior Member since 2006. In May 2009 he started his role as Director of the DVB Project Office in Geneva, Switzerland. From November 2008 he was with Albis Technologies Ltd, the former R&D Division of Siemens Switzerland. He joined Siemens Switzerland Ltd in 2001 in the area of system design of next generation systems, combining SDH and packet oriented transmission schemes. Since 2003, he was instrumental in the end-to-end video system design and STB architecture for the Siemens Home Entertainment IPTV Solution. From 1995 to 2001 he was with SES-ASTRA, a European Satellite operator based in Luxembourg, where he was involved the introduction of digital Television based on DVB-S and in planning and development of the ASTRA Broadband Interactive system. In this role he actively participated in ETSI and DVB working groups and has been the Rapporteur for several standards documents. Prior to SES-ASTRA, from 1990 to 1994, he was leading a team of product development engineers in the area of professional audio and video transmission over telecomm networks at Philips Kommunikations Industrie (PKI) AG in Nurnberg, Germany.

19.2.2 BART BRUSSE

Bart, who has a bachelor's degree in sociology and a master's degree in mass communication from the University of Nijmegen, Netherlands, has worked for many years as Head of Corporate Strategy and Business Development for Casema (now Ziggo), one of the major Dutch cable operators. Since 2000 he has been active as an independent consultant and has worked on various projects for many clients in the European and global broadcast and digital media industry such as UPC, KDG, Viaccess and Nielsen Media Research. During the last years, Bart has managed projects in the 6th and 7th Framework Programme of the European Commission. Currently, he is managing the ReDeSign project and is active in DVB as the Chairman of the DVB-CM-C2 group.

19.2.3 Christoph Schaaf

Christoph studied electrical engineering at the University in Darmstadt, Germany. In 1981 he joined Deutsche Telekom Research Institute working on digital multiplexing technologies. He was involved in the DVB-C development in 1994. Since 1996 he headed several teams developing a digital cable platform for Deutsche Telekom cable. In 2002 he joined Kabel Deutschland, when Deutsche Telekom sold the CATV business. 2007 he was one of the initiators of the DVB-C2 project and was appointed as the DVB-C2 project leader.

19.2.4 Frank Hermann

Frank received his diploma for communication technologies from the Technical University of Braunschweig. From 1994 he was involved in the Eureka 147 Project developing the first digital broadcasting system in Europe – DAB. In 1996 Frank joined the Panasonic R&D Center Germany and still put emphasis for a few more years on DAB – chairing WorldDAB's Technical Committee at the time when DMB and DAB+ were created. In 2006 he became project leader for the DVB projects of the Panasonic Frankfurt labs. In DVB TM-T2 he chaired the working group that developed the system layer part of the T2 standard and has also contributed to the development of DVB-C2.

19.2.5 Marten Kabutz

Marten received B.Sc. and M.Sc. degrees in electrical engineering from the University of Cape Town in 1991 and 1995, respectively. In 1996, he joined Thomson and was lab manager for demodulator architectures until December 2009. In 2010, Marten co-founded Ubiso, an IP-design company specializing in error correction and signal processing IP-cores for integrated circuits and FPGAs. The Ubiso team, an official Thomson spin-off, has been at the forefront of innovation in Forward Error Correction (FEC) inside Thomson, from the beginning of error control coding in CD/DVD to advanced iterative techniques like soft-decision codes for DVB standards. Marten and his team have actively contributed to standardization bodies (e.g. DVB-C2) and developed IC architectures for optical disc (BluRay, DVD/CD) and transmission standards (Mobile-ATSC, DTMB and DVB).

19.2.6 Jörg Robert

Jörg studied electrical engineering at Ilmenau Technical University and Braunschweig Technical University, Germany. His Diploma thesis was on the subject of impulsive noise cancellation algorithms for DVB-T. Since January 2006 he has worked as researcher at the Institute for Communications Technology of Braunschweig Technical University, Germany. He was member of DVB TM-T2 and DVB TM-C2 , in which he chaired the subgroup on framing structure and synchronization. Currently, he is writing his PhD thesis on the application of MIMO in broadcasting networks.

19.2.7 Woo Suk Ko (Ph. D.)

Woo-Suk received his B.S and M.S degree in Electrical and Electronic Engineering and Ph.D. in digital signal processing from Yonsei University, Seoul Korea. Since 2002 he has been working for Digital TV Laboratory, LG Electronics as a chief research engineer and developed several demodulators including Terrestrial / Satellite DMB and DVB-T. He has research interests in digital communication and broadcasting system and contributed technologies for DVB-T2 and C2 standardization.

19.2.8 Dr. Sam Atungsiri

Sam received a BSc in Computer Systems Engineering from the University of Wales in 1987, and MSc (1988) and PhD (1991) both in digital communications from the University of Surrey. Dr Atungsiri is currently a Senior Consultant design engineer at Sony's Semiconductor

Development Centre in Basingstoke in the United Kingdom. His main specialism is in the physical layer issues of communication and broadcast systems. He led Sony's contribution to the DVB technical committees on DVB-H and DVB-T2. He has also participated in DVB-C2 helping to formulate Sony's C2 strategy during the standards process, contributing technology and drafting some sections of the specification and implementation guidelines.

19.2.9 Taku Yamagata

Taku received his Bachelor and Master of Science degrees from Shizuoka University in Japan in 1992 and 1994, respectively. From 1994 to 2006, he was with Sony Corporation where he was responsible for research and development of algorithms and hardware architectures for various modem projects including DVB-C, DVB-T/H, DVB-S, ISDB-S, ISDB-T/T1seg and 802.11a/b/g. In March 2006, he joined Imagination Technologies, UK, where he is currently a Senior Engineer leading a multi-standard cable TV demodulation project as part of Imagination Technologies' programmable multi-standard solution for both broadcast and communications.

19.2.10 Lothar Stadelmeier

Lothar studied electrical engineering at University of Stuttgart, Germany. In 1997 he joined BetaResearch, being responsible for DVB frontend integration in set-top boxes. Since 1999 he has been with Sony, working on different digital communication systems and standardization groups such as DVB, Powerline Communications, and Wireless Networks.

19.2.11 Sang-Chul Moon

Sang works as Senior Research Engineer at CNC Gr., Digital TV Lab. of LG Electronics Inc. He holds a M.S. in Electrical and Electronic Engineering of KAIST, Korea (completed in 2003) and a B.S. in Electronic Engineering from Hanyang University, Korea (completed in 2001). Sang worked in various R&D environments where he gained experience in technologies such as T-DMB receiver baseband development and DVB-T receiver baseband development. He was actively involved in the development of DVB-C2 specification during 2008 and 2009 and DVB-T2 specification during 2007 and 2008 where he mad various technical contributions.

19.2.12 Seho Myung (Ph. D.)

Seho received the B.S. degree in electronics engineering from Korea Advanced Institute of Science and Technology (KAIST) in 2001 and the M.S. and PH.D. degrees in electronics engineering from the Pohang University of Science and Technology (POSTECH), Korea, in 2003 and 2006, respectively. From March 2006 to March 2007, he was a postdoctoral research fellow in the Department of Electronics and Electrical Engineering, POSTECH. Since April 2007, he has been a senior engineer at Samsung Electronics Co. He is working at Multimedia Global Standards Group in Digital Media & Communications R&D Center.

19.2.13 Michael Heisenberg

In 1983 Michael received a Dipl.-Ing. (Master in Electronic Engineering) from the Technical University of Munich. After joining Kathrein-Werke, he was involved in the development of professional CATV head-end equipment. Since 1989 he has been R&D Manager for the Headend business segment. In 1998, Michael was appointed as leader of the SMATV development at Kathrein. Since 2007, he has been Senior Manager New Technologies focusing on digital signal processing. He represents Kathrein in several international bodies.

19.2.14 Philipp Hasse

Philipp studied Computer and Communications Systems Engineering at Technische Universität Braunschweig, Germany. His Diploma thesis was on "Data Modeling of an Engine

Test Bench". Since July 2008, he has been working as a research associate at the Institut für Nachrichtentechnik of Technische Universität Braunschweig, Germany. Currently, his scope of duties comprises the work for the ReDeSign project, which includes the cooperation with the DVB project regarding work on DVB-C2, in particular the chairmanship of the DVB-C2 Validation & Verification group.

19.2.15 Dr. Dirk Jaeger

Dirk is working with the Institut für Nachrichtentechnik (Communications Technology) at Technische Universität Braunschweig where he has been involved in various research and project management activities. He coordinates, for instance, the research project ReDeSign investigating next generation cable technologies. He is active as Chairman of technical committees of cable standardization bodies (e.g. CENELEC TC 209) and associations (e.g. ITG-FA 3.3) and has been appointed into various advisory committees. Dirk holds a diploma in Communications Technology from Technische Universität Braunschweig where he also achieved a Doctorate in Engineering in June 1998. He is member of FKTG (D) and SCTE (UK).

19.2.16 Jan De Nijs (Ph.D.)

Jan de Nijs is currently working as senior consultant and senior scientist in the area of cable networks and infrastructures. He has a Masters degree in Theoretical Physics (1985) and a Ph.D. degree in Applied Physics (1989), both from the University of Twente in The Netherlands. After an academic career in silicon technology, he joined the group of Wireless Communications of the Technical University of Delft in 1998. In 2001, he started working at TNO Information and Communication Technologies as a senior consultant/scientist in the area of access networks with a focus on cable and fixed wireless networks. He is charged with the development of a knowledge portfolio on these topics.

19.2.17 Ernst Freese

Ernst studied communication and information engineering at the University of Applied Sciences Bielefeld and University (GH) Siegen, Germany. In 1981 he joined FUBA, Germany, as an R&D Engineer and was assigned to Director R&D in 1996. Since 1998, he continued in this role within FUBA successor companies General Instrument and Motorola BCS, Bad Salzdetfurth, Germany. In 2002, Ernst acted as General Manager of Motorola BCS. In October of this year, he co-founded BLANKOM Digital GmbH. Since then he has been acting as Managing Partner for that company. Ernst is Chairman of the German standardisation committee DKE K 735: “Cable networks and antennas for television signals, sound signals and interactive services”. He is also member of the Technical Council for Satellite and Cable within ZVEI, Germany, member of DVB Technical Module and of FKTG, Germany.

19.2.18 Peter Flötgen

Peter received his diploma degree in Information Technologies from the University of Duisburg. He has been working in the cable TV business since 1991 and stated his carrier in operations and customer service of Deutsche Telekom in North Rhine-Westphalia. Since 2003 he has worked in the engineering department of Unitymedia (formerly ish). Today, Peter is responsible for the technical product development TV services, which includes digital TV headend equipment as well as set-top box technologies.

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19.4 Acronyms and Abbreviations

ADC	analogue to Digital Conversion / Analogue to Digital Converter
ASIC	Application Specific Integration Circuit
aTV / ATV	analogue Television
BB	Baseband
BCH	Bose Chaudhuri Hoquenghem
BER	Bit Error Rate
CA	Conditional Access
CENELEC	Comité Européen de Normalisation Electrotechnique / European Committee for Electrotechnical Standardization
CM	Cable Modem
CMTS	Cable Modem Termination System
CINR	Carrie-to-Intermodulation-Noise Ratio
CNR	Carrier-to-Noise Ratio
COFDM	Coded OFDM

CPE	Common Phase Error
CPE	Customer Premises Equipment
CSO	Composite Second Order
CTB	Composite Triple Beat
DAC	Digital to analogue Conversion / Digital to Analogue Converter
Davic	Digital Audio Visual Council
dB	Decibel
DOCSIS	Data over Cable Service Interface Specification
dRadio	Digital Radio
dTV / DTV	Digital Television
DVB	Digital Video Broadcasting
EN	European Norm
Eq	Equation
ETSI	European Telecommunications Standards Institute
EU	European Union
FEC	Forward Error Correction
FM	Frequency Modulation
FPGA	Field Progammable Gate Array
Gbps	Gigabit per Second
GS	Generic Stream
GSE	Generic Stream Encapsulation
HD / HDTV	High Definition Television
HFC	Hybrid Coax Fibre
HSI	High Speed Internet
ICI	Inter-Carrier Interference
IEC	International Electrotechnical Commission
IM	Intermodulation
IP	Internet Protocol
LDPC	Low Density Parity Check
M-CMTS	Modular CMTS
MBaud	Mega Baud
Mbps	Megabit per Second
MHz	Megahertz
MPEG	Moving Picture Experts Group
ms	Millisecond
MSO	Mean Opinion Score
Msps	Mega Symbols per Second
OFDM	Orthogonal Frequency Division Multiplex
PAL	Phase Alternating Line
PAPR	Peak to Average Power Ratio
PDF	Probability Density Function
PHY	Physical Layer
PLP	Physical Layer Pipe
Px	Power level of carrier x
QAM	Quadrature Amplitude Modulation
QEF	Quasi Error Free
QoS	Quality of Services
QPSK	Quadrature Phase Shift Keying
RF	Radio Frequency

SCTE	Society of Cable Telecommunications Engineers renamed to Society for Broadband Professionals
SECAM	Séquentiel Couleur à Mémoire
SD / SDTV	Standard Definition Television
SNR	Signal-to-Noise Ratio
SoC	System on Chips
TI	Time Interleaver
TS	Transport Stream
TV	Television
VCM	variable coding and modulation
VOD	Video on Demand